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(54) RADIO RECEIVING DEVICE AND RADIO RECEIVING METHOD

(57) A reception weight calculation section 203 calculates reception weights W_1 and W_2 every antenna using an optimal directional control method in order to improve interference cancellation effect, an arrival direction estimation section 204 estimates a direction of arrive of a received signal for each antenna to calculate steering vectors S_1 and S_2 for each antenna, and a replica weight calculation section 211 calculates replica weights W_{r1} and W_{r2} using the reception weights W_1, W_2 and steering vectors S_1, S_2.

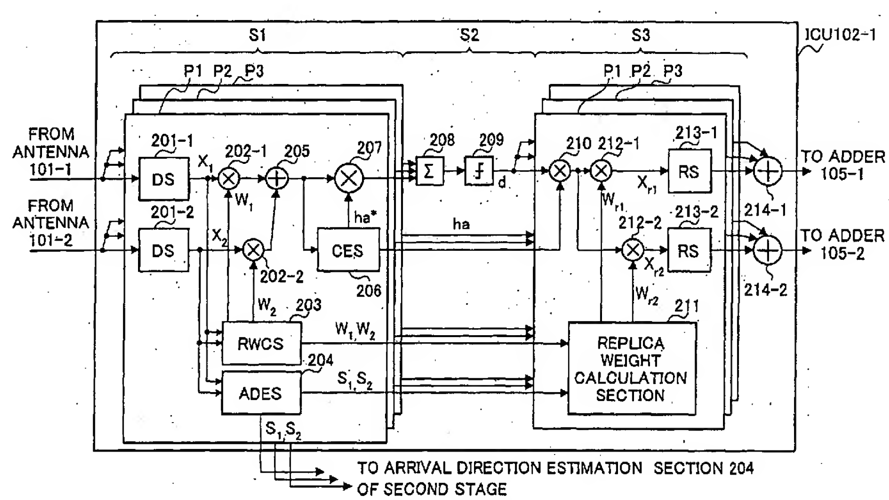

FIG.2

201 DS : DESPREADING SECTION
203 RWCS : RECEPTION WEIGHT CALCULATION SECTION
204 ADES : ARRIVAL DIRECTION ESTIMATION SECTION
206 CES : CHANNEL ESTIMATION SECTION
213 RS : RE-SPREADING SECTION

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Description

Technical Field

[0001] The present invention relates to a radio receiving apparatus and a radio receiving method used in a mobile communication system of CDMA (Code Division Multiple Access).

Background Art

[0002] In a mobile communication system of CDMA, since a plurality of user signals is transmitted in the same band, a signal that a radio receiving apparatus receives is subjected to interference by various signals to cause deterioration of characteristics.

[0003] An array antenna is known as an apparatus for eliminating the interference. The array antenna is composed of a plurality of antenna elements, and is capable of setting reception directivity freely by providing adjustment of each of amplitude and phase to a signal received by each antenna element. In this case, adjustment of amplitude and phase provided to the received signal can be carried out by multiplying the received signal by weighting factor (hereinafter referred to as "reception weight"). The radio receiving apparatus can intensively receive only a signal coming from a desired direction by adjusting the reception weight by which the received signal is multiplied.

[0004] Another apparatus for eliminating the interference, an interference canceller is known. The interference canceller is a technique for canceling a signal (interference) transmitted from other communication partners other than current communication partner from the received signal to extract a desired signal from the received signal. Conventionally, as an apparatus for canceling interference signals, there are apparatuses described in 1) "Sequential Channel Estimation Type Serial Canceller Using a pilot Symbol in DS-CDMA (Technical Bulletin, RCS95-50, July, 1995, Radio Communication System Research Society of the Institute of Electronics, Information and Communication Engineers)" authored by Sawahashi, Miki, Andoh, and Higuchi, 2) "Sequential Transmission Line Estimation Type CDMA Multistage Interference Canceller Utilizing a Symbol Replica Process (Technical Bulletin, RCS96-171, February, 1997, Radio Communication System Research Society of the Institute of Electronics, Information and Communication Engineers)" authored by Yoshida and Ushirokawa, and 3) "Study of CDMA Interference Canceller in an Upstream Line(Technical Bulletin, RCS96-121, January, 1997, Radio Communication System Research Society of the Institute of Electronics, Information and Communication Engineers)" written by Uesugi, Katoh, and Honma. The above three apparatuses are hereinafter referred to as 1) a serial type interference canceller, 2) a parallel type interference canceller, and 3) a symbol ranking type canceller.

[0005] Here, it can be expected that the use of combination of the array antenna and the interference canceller provide a larger interference cancellation effect than each independent use.

[0006] However, in the radio communication system that provides reception directivity to each channel corresponding to each communication partner by use of the array antenna, degree of interference with respect to each communication partner is different from one communication partner to another. Accordingly, in the case of applying the interference canceller to such the system, it is necessary to individually provide the interference canceller to each channel corresponding to each communication partner. Hence, the simple combination of the array antenna and the interference canceller increases the amount of calculations and the apparatus scale, making it difficult to implement such an apparatus in consideration given to actual hardware, design.

Disclosure of Invention

[0007] It is an object of the present invention is to provide a radio receiving apparatus and a radio receiving method that are capable of receiving a desired signal with high quality in an apparatus of small scale without providing an interference canceller to each channel corresponding to each communication partner even when the array antenna and the interference canceller are combined.

[0008] In order to attain the above object, the present invention generates a replica signal every signal received by each antenna of the array antenna to make it possible to receive a desired signal with high quality in an apparatus of small scale without providing an interference canceller to each channel corresponding to each communication partner even when the array antenna and the interference canceller are combined.

[0009] Particularly, the present invention is characterized in that a reception weight by which an optimal radiation pattern is formed is calculated to improve an interference cancellation effect without limiting to a calculation algorithm of the reception weight. Moreover, the present invention is characterized in that the reception weight is sequentially updated using a signal from which a interference signal is sequentially eliminated to sequentially generate a radiation pattern with high reliability, whereby further improving the interference cancellation effect.

Brief Description of Drawings

[0010]

FIG. 1 is a block diagram of a main part illustrating a schematic configuration of a radio receiving apparatus according to Embodiment 1 of the present invention;

FIG. 2, is a block diagram of a main part illustrating

a schematic configuration of ICU of each of first and second stages of an interference signal canceling apparatus according to Embodiment 1 of the present invention;

FIG. 3 is a block diagram of a main part illustrating a schematic configuration of ICU of a third stage of the interference signal canceling apparatus according to Embodiment 1 of the present invention;

FIG. 4 is a view of a radiation pattern formed by a beam steering.

FIG. 5 is a view of a radiation pattern formed by a null steering.

FIG. 6 is a block diagram of a main part illustrating a schematic configuration of a radio receiving apparatus according to Embodiment 2 of the present invention;

FIG. 7A is a view showing one example of a radiation pattern formed by each ICU of the radio receiving apparatus according to Embodiment 2 of the present invention;

FIG. 7B is a view showing one example of a radiation pattern formed by each ICU of the radio receiving apparatus according to Embodiment 2 of the present invention; and

FIG. 7C is a view showing one example of a radiation pattern formed by each ICU of the radio receiving apparatus according to Embodiment 2 of the present invention.

Best Mode for Carrying Out the Invention

[0011] Embodiments of the present invention will be specifically described with reference to the drawings accompanying herewith.

(Embodiment 1)

[0012] FIG. 1 is a block diagram of a main part illustrating a schematic configuration of a radio receiving apparatus according to Embodiment 1 of the present invention. The following will explain the case in which the number of stages of the interference canceller is 3, the number of communication partners is 3, and the number of multipaths is 3. It is noted that these numbers are just one example, and this Embodiment is not limited to these number.

[0013] In addition, as illustrated in FIG. 1, since the first stage and the second stage have the same configuration, the same reference numerals are added to the same structural parts, and the explanation of the second stage is omitted.

[0014] In FIG. 1, a signal received via an antenna 101-1 is inputted to ICUs (Interference Canceling Units) 102-1 to 102-3 and a delayer 103-1 provided to correspond to the antenna 101-1. Similarly, a signal received via an antenna 101-2 is inputted to ICUs 102-1 to 102-3 and a delayer 103-2 provided to correspond to the antenna 101-2.

[0015] ICUs 102-1 to 102-3 are provided to correspond to communication partners 1 to 3, respectively, and each generates a replica signal in connection with each of the signals received via the antennas 101-1 and 101-2. The replica signals generated by the ICUs 102-1 to 102-3 are inputted to adders 104-1 and 104-2 provided to correspond to the antennas 101-1 and 101-2, and are inputted to adders 105-1 and 105-2. The configuration of each of the ICUs 102-1 to 102-3 will be described later.

[0016] The delayers 103-1 and 103-2 delay the received signals by processing time of ICUs 102-1 to 102-3, and each outputs the resultant to each of the adders 104-1 and 104-2.

[0017] In the adder 104-1, the replica signals of communication partners 1 to 3 for the signal received via the antenna 101-1 are subtracted from the signal received via the antenna 101-1. Also, in the adder 104-2, the replica signals of communication partners 1 to 3 for the signal received via the antenna 101-2 are subtracted from the signal received via the antenna 101-2. This eliminates the replica signals of all communication partners from the signals received via the respective antennas. Signals (residual signals) obtained by eliminating the replica signals of all communication partners from the received signals are inputted to the adders 105-1 and 105-2, respectively, and are inputted to the delayers 103-1 and 103-2 of the second stage.

[0018] In the adders 105-1 and 105-2, the replica signals for the signals received via the antennas 101-1 and 101-2 and the residual signals are added every communication partner. This eliminates the replica signal of communication partner 1, the replica signal of communication partner 2, the replica signal of communication partner 3 from the received signals every antenna. Namely, when attention is paid to communication partner 1, the signal from communication partner 2 and the signal from communication partner 3, which cause interference with communication partner 1, are eliminated from the received signal to obtain a desired signal for communication partner 1 every antenna. The similar processing is carried out, so that the signals of other communication partners causing interference are eliminated from the received signals, so that the desired signal for communication partner 2 and the desired signal for communication partner 3 can be obtained every antenna. The obtained desired signals are inputted to ICUs 102-1 to 102-3 of the second stage, respectively.

[0019] According to the radio receiving apparatus of this embodiment, the same processing as performed in the first stage is repeated in the second stage, so that the accuracy of replica signal is improved and that of the interference signal cancellation is improved. In other words, the more the number of stages are increased, the more the inference signals sent from the other communication partners that cause interference with the respective communication partners are eliminated.

[0020] The signals added by the adders 105-1 and

105-2 of the second stage are inputted to ICUs 106-1 to 106-3 of the third stage, and are demodulated. This obtains demodulated signals 1 to 3 of the communication partners 1 to 3. The configuration of each of the ICUs 106-1 to 106-3 will be described later.

[0021] An explanation will be next given of ICUs 102-1 to 102-3 and ICUs 106-1 to 106-3. FIG. 2 is a block diagram of a main part illustrating a schematic configuration of ICU of each of first and second stages of an interference signal canceling apparatus according to Embodiment 1 of the present invention. Also, FIG. 3 is a block diagram of a main part illustrating a schematic configuration of ICU of a third stage of the interference signal canceling apparatus according to Embodiment 1 of the present invention. Additionally, ICUs 102-1 to 102-3 of the first and second stages have the same configuration and operation, respectively. Also, ICUs 106-1 to 106-3 of the third stage have the same configuration and operation. Accordingly, in the explanation set forth below, the ICU 102-1 of the first stage and the ICU 106-1 of the third stage corresponding to the communication partner 1 are explained, and the explanation of the respective ICUs corresponding to the communication partner 2 and the communication partner 3 is omitted. Moreover, the ICU 102-1 shown in FIG. 2 and the ICU 106-1 shown in FIG. 3 are configured on the assumption that the number of multipath to the radio receiving apparatus is 3. In FIGS. 2 and 3, the respective configuration parts for the respective paths are shown by P1 to P3, respectively. Since the respective configuration parts for the respective paths have the same configuration and operation, only the first path P1 is illustrated, and the explanation of the second path P2 and third path P3 is omitted.

[0022] In FIG. 2, the ICU 102-1 briefly includes a preceding stage S1 in which the signals received by the respective antennas 101-1 and 101-2 are subjected to despreading and then the resultants are multiplied by reception weights of the receptive antennas, respectively, an intermediate stage S2 in which RAKE combining and provisional decision are carried out, and the last stage S3 in which the signal subjected to provisional decision is multiplied by a weighting factor for generating a replica signal (hereinafter referred to as "replica weight") to generate a replica signal.

[0023] The signal received via the antenna 101-1 is despread by a despreading section 201-1 and the signal received via the antenna 101-2 is despread by a despreading section 201-2. Despread signals X_1 and X_2 are inputted to multipliers 202-1, 202-2, a reception weight calculation section 203, and an arrival direction estimation section 204.

[0024] The reception weight calculation section 203 calculates weights W_1 and W_2 of each antenna, and outputs the resultants to multipliers 202-1 and 202-2, and a replica weight calculation section 211. Since the reception weight calculation section 203 is provided every path and every communication partner, making it possible to calculate the reception weights each being different every path and every user. The calculation method for the reception weight will be described later.

[0025] The arrival direction estimation section 204 estimates a direction of arrival of the received signal every antenna, and outputs steering vectors S_1 and S_2 of the respective antennas to the replica weight calculation section 211, and the arrival direction estimation section 204 of the second stage. Here, the reason why the arrival direction estimation section 204 of the first stage outputs the steering vectors S_1 and S_2 to the arrival direction estimation section 204 of the second stage is as follows. Specifically, the arrival direction estimation section 204 of the second stage averages the steering victors calculated in the first stage and the steering victors calculated in the second stage every path, and uses the resultant as a steering vector in the second stage. This makes it possible to increase the accuracy of the steering vector as the operation goes to the last stage. In other words, the accuracy of the direction of arrival can be improved as the operation goes to the last stage, making it possible to improve the accuracy of the calculation of the replica weight.

[0026] Here, since the signal inputted to each stage is a signal from which an interference signal is eliminated in the previous stage, the signal whose interference state changes every stage is inputted. Hence, according to this embodiment, the reception weight calculation section 203 and the arrival direction estimation section 204 are provided on a stage-by-stage basis. This makes it possible to adaptively change the radiation pattern in accordance with the state of the inference signal at this point on the stage-by-stage basis. Hence, according to this embodiment, the radiation pattern and the replica signal can be accurately generated. This eliminates the useless processing wherein interference cancellation using directional control is further performed to interference that can be sufficiently cancelled by only interference cancellation processing, conversely; interference cancellation processing is further performed to interference that can be sufficiently cancelled by only directional control.

[0027] Moreover, according to this embodiment, the direction of arrival of the signal from which the interference signals are sequentially cancelled is estimated. Hence, the accuracy of estimation of the direction of arrival is improved as the operation goes to the last stage. Accordingly, since the interference cancellation having good performance can be carried out with a relatively small number of stages, the apparatus scale can be reduced.

[0028] Despread signals X_1 and X_2 are multiplied by reception weights W_1 and W_2 by the multipliers 202-1 and 202-2, respectively, and the resultant is added by an adder 205. This carries out array combining. The signal subjected to array combining is outputted to a channel estimation section 206 and is outputted to a multiplier 207.

[0029] The channel estimation section 206 performs the channel estimation based on the signal subjected to the array combining, and outputs the resultant to a complex conjugate h_a* of a channel estimation value h_a to the multiplier 207, and outputs the channel estimation value h_a to a multiplier 210. The multiplier 207 multiplies the signal subjected to the array combining by the complex conjugate h_a* of the channel estimation value. This compensates for phase rotation of the signal subjected to the array combining.

[0030] The signal, which has been subjected to the array combining of each of paths P1 to P3 and which has been multiplied by the complex conjugate h_a* of the channel estimation value, is subjected to RAKE combining by an adder 208 of the intermediate stage S2. The result obtained by RAKE combining is temporarily decided by a decider 209. A signal d subjected to temporarily decision is multiplied by the channel estimation value h_a by a multiplier 210 for each of paths P1 to P3, and the resultant is inputted to multipliers 212-1 and 212-2, respectively.

[0031] A replica weight calculation section 211 calculates replica weights W_{r1} and W_{r2} using reception weights W_1 and W_2 and steering vectors S_1 and S_2, and outputs the resultant to the multipliers 212-1 and 212-2, respectively. The method for calculating the replica weight will be described later.

[0032] The multipliers 212-1 and 212-2 multiply the signals outputted from the multiplier 210 by replica weights W_{r1} and W_{r2}, respectively. This obtains replica signals X_{r1} and X_{r2} corresponding to X_1 and X_2, respectively. The replica signals Xr_1 and Xr_2 are spread by re-spreading sections 213-1 and 213-2, respectively, and the resultants are inputted to adders 214-1 and 214-2. The replica signals Xr_1 and Xr_2 re-spread for each of paths P1 to P3 are added by adders 214-1 and 214-2, respectively, and the resultants are inputted to adders 105-1 and 105-2.

[0033] Next, the ICU 106-1 of the third stage will be described. As illustrated in FIG. 3, the ICU 106-1 of the third stage has substantially the same structure as that of the preceding stage S1 and that of the intermediate stage S2 of the ICU 102-1 of FIG. 2. Accordingly, the same reference numerals are added to the same configuration parts as those of the ICU102-1 of FIG. 2, and the explanation of the ICU 106-1 of the third stage will be omitted. The ICU 106-1 is different from the ICU 102-1 in the point that there is no the arrival direction estimation section 204 provided in the ICU 102-1. This is because in the third stage, demodulated signal 1 is outputted instead of the replica signal, and therefore the replica weight necessary for generating the replica signals not required, whereby steering vector necessary for calculating the replica weight is not required also.

[0034] An explanation will be next given of the method for calculating the reception weights W_1 and W_2, and the method for calculating the replica weights W_{r1} and W_{r2}.

[0035] The method of directional control using the array antenna is largely divided into directional control carried out by a beam steering and directional control carried out by a null steering.

[0036] The beam steering is a method in which interference from the other communication partners is eliminated by generating such a radiation pattern that directs directivity to a direction where a desired communication partner exists. On the other hand, the null steering is a method in which interference from the other communication partners is eliminated by generating a radiation pattern that forms a null point in a direction where a desired communication partner exists.

[0037] In the case of performing array reception using the beam steering, the signals received by the respective antennas are multiplied by in-phase addition weights as reception weights W_1 and W_2 such that the signals received by the respective antennas are added in a state that they all are in phase with each other. Here, the in-phase addition weights are weights that adjust only phases of the signals received by the respective antenna. For this reason, in the case of using the in-phase addition weights as reception weights W_1 and W_2, the signals subjected to provisional decision are multiplied by complex conjugates of reception weights W_1 and W_2 as replica weights Wr_1 and Wr_2 in order to return the adjusted phases to the original. This makes it possible to generate replica signals Xr_1 and Xr_2 for each antenna.

[0038] However, in the case of the beam steering (namely, in-phase addition weight), the radiation pattern is not in a pointed form as illustrated in FIG. 4. Moreover, in the case of the beam steering, control is performed in such a way that the center of the radiation pattern is directed to the direction where a desired communication partner exists. For this reason, when the direction where the desired communication partner exists and the direction where the communication partner, which causes interference, exists are close to each other or when transmission power of the communication partner, which causes interference, is greater than that of the desired communication partner, it is impossible to sufficiently eliminate interference with respect to the desired communication partner.

[0039] Morespecifically, as illustrated in FIG. 4, when communication partner 2 exists closely in the direction where a desired communication partner 1 exists, the signal sent from the communication partner 2 that causes interference with the communication partner 1 cannot be fully eliminated in the case of the beam steering. For this reason, the gain of the desired communication partner 1 becomes extremely small as compared with the case in which there is no interference from the communication partner 2.

[0040] On the other hand, in the case of the null steering, such a radiation pattern that directs the null point to the direction, where the communication partner 2 that causes interference exists, is formed in connection with the desired communication partner 1 as illustrated in

FIG. 5. This makes it possible to fully eliminate the signal sent from the communication partner 2 that causes interference with the communication partner 1. As a result, the gain of the desired communication partner 1 becomes extremely large as compared with the case of using the beam steering. In this way, it is useful to perform the array reception using the null steering at the time of eliminating the interference signal.

[0041] Accordingly, the radio receiving apparatus of this embodiment performs the array reception using the null steering. In other words, the reception weight calculation section 203 shown in FIG. 2 calculates reception weights W_1 and W_2 by a control algorithm using, for example, MMSE (Minimum Means Square Error) as a code so as to obtain the null point.

[0042] However, reception weights W_1 and W_2 thus obtained are not the weights that adjust only the phases of the signals received by the respective antennas. Hence, in the case of performing the array reception using the null steering, replica signals X_{r1} and X_{r2} for every antenna cannot be generated by multiplying the signals subjected to provisional decision by complex conjugates of reception weights W_1 and W_2 as replica weights W_{r1} and W_{r2}.

[0043] For this reason, according to this embodiment, the replica weight calculation section 211 shown in FIG. 2 calculates a replica weight W_{rk} in the following way. Additionally, in this embodiment, since the number of array antennas is two, k is 1 or 2.

[0044] It is assumed that a signal subjected to provisional decision by the decider 209 is d, a steering vector of each antenna obtained by the arrival direction estimation section 204 is S_k and a channel estimation value of a signal X_k received by each antenna is h. The replica signal X_{rk} can be expressed by the following equation (1):

$$X_{rk} = dhS_k \quad (1)$$

[0045] Additionally, since it is assumed that fading correlation between the array antennas is 1, the channel estimation values of the signals received by the respective antennas are all h.

[0046] Moreover, it is assumed that the channel estimation value of the signal subjected to array combining obtained by the channel estimation section 206 is h_a and a reception weight by which the signal X_k received by each antenna is multiplied is W_k. The following equation is established.

$$dh_a = \sum_{k=1}^{n} X_{rk} W_k \quad \cdots (2)$$

where n denotes the number of antenna.

[0047] Substitution of equation (1) into equation (2) yields the following equation (3):

$$dh_a = \sum_{k=1}^{n} dhS_k W_k \quad \cdots (3)$$

[0048] From the equation (3), the following equation (4) is established:

$$h = \frac{h_a}{\sum_{k=1}^{n} S_k W_k} \quad \cdots (4)$$

[0049] Next, substitution of equation (4) into equation (1) yields the following equation (5):

$$X_{rk} = \frac{dS_k h_a}{\sum_{k=1}^{n} S_k W_k} \quad \cdots (5)$$

[0050] Moreover, the replica signal X_{rk} can be expressed by the following equation (6):

$$X_{rk}=dh_a W_{rk} \quad (6)$$

[0051] Then, comparison between equation (5) and (6) is performed and the following equation (7) can be obtained as a replica weight W_{rk} by the replica weight calculation section 211.

$$W_{rk} = \frac{S_k}{\sum_{k=1}^{n} S_k W_k} \quad \cdots (7)$$

[0052] Accordingly, the radio receiving apparatus of this embodiment can calculate the replica weight W_{rk} without limitation of the kinds of the reception weights even if any kind of reception weight is used as a reception weight W_k.

[0053] Therefore, the radio receiving apparatus of this embodiment can generate the replica signal X_{rk} every antenna even if the replica weight W_{rk} is not the complex conjugate of the reception weight W_k. In other words, since the kind of reception weight used in the radio receiving apparatus of this embodiment is not limited to the in-phase addition weight, the radio receiving appa-

ratus of this embodiment can perform the array reception using the null steering having high interference cancellation effect.

[0054] The above has explained the case in which the array reception is performed using the null steering as one example. The radio receiving apparatus of this embodiment can generate the replica signal even if any kind of reception weight is used, so that the method of the array reception is not limited to the null steering.

[0055] For example, in the case where the radio receiving apparatus of this embodiment performs the array reception using the beam steering, the arrival direction estimation section 204 outputs the steering vector S_k to the reception weight calculation section 203, and the reception weight calculation section 203 calculates the reception weight W_k as a complex conjugate S_k^* of the steering vector S_k.

[0056] Namely, W_k in the above equation (7) is equal to S_k^*.

$$W_k=S_k^* \tag{8}$$

[0057] Accordingly, the replica weight calculation section 211 calculates the replica weight W_{rk} using the above equation (7) to obtain the following equation (9):

$$W_{rk}=S_k \tag{9}$$

[0058] Accordingly, since the replica weight W_{rk} serves as a complex conjugate of the reception weight W_k, the radio receiving apparatus of this embodiment can use the in-phase addition weight also as a reception weight W_k.

[0059] In this way, according to the radio receiving apparatus and the radio receiving method of this embodiment, the array reception is performed using the optimal directional control method in order to improve the interference cancellation effect without limiting to a calculation algorithm of the reception weight, and the replica signal can be generated every signal received by each antenna of the array antenna. This makes it possible to receive a desired signal with high quality in an apparatus of small scale even when the array antenna and the interference canceller are combined.

[0060] Moreover, according to the radio receiving apparatus and the radio receiving method of this embodiment, the reception weight can be updated in accordance with the change in the state of interference. This makes it possible to generate the radiation pattern and the replica signal accurately.
Accordingly, according to the radio receiving apparatus and the radio receiving method of this embodiment,' since the interference cancellation having good performance can be carried out with a relatively small number of stages, the apparatus scale can be reduced.

[0061] Still' moreover, according to the radio receiving apparatus and the radio receiving method of this embodiment, since, the accuracy of the estimation of the direction of arrival can be improved as the operation goes to the last stage, the accuracy of the calculation of the replica weight can be improved.

(Embodiment 2)

[0062] The radio receiving apparatus and the radio receiving method of this embodiment are to eliminate the interference signals sequentially every communication partner in one stage and to update the reception weights sequentially every communication partner in one stage.

[0063] FIG. 6 is a block diagram of a main part illustrating a schematic configuration of a radio receiving apparatus according to Embodiment 2 of the present invention. Additionally, ICUs 606-1 to 606-3 shown in FIG. 6 have the same configuration as that of the ICU 102-1 shown in FIG. 2, and the detailed explanation of each ICU is omitted. It is noted that the ICU 606-1 and 606-2 of the third stage shown in FIG. 6 adopt the configuration that output the replica signal and output demodulated signal 1 and 2, respectively. Also, the ICU 606-3 of the third stage shown in FIG. 6 adopts the same configuration as that of the ICU 106-1 shown in FIG. 3 so as to output a demodulated signal'3.

[0064] In addition, as illustrated in FIG. 6, since the first to third stages have the same configuration, the same reference numerals are added to the same structural parts, and the explanation of the second and third stages are omitted.

[0065] Signals received via antennas 601-1 and 601-2 are inputted to delayers 602-1,602-2, and 603-1, 603-2, respectively. The received signals inputted to the delayers 602-1,602-2 are delayed by a given time and outputted to the second stage. The received signals inputted to the delayers 603-1,603-2 are delayed by a given time, and outputted to the ICU 606-1 and outputted to delayers 604-1 and 604-2.

[0066] In the ICU 606-1, a reception weight, a steering vector, and a replica signal of the communication partner 1 are generated every antenna based on the received signal. The replica signal of the communication partner 1 generated every antenna is inputted to each of adders 607-1 and 607-2, and the steering vector every antenna is inputted to the ICU 606-1 of the second stage.

[0067] In the adders 607-1 and 607-2 connected to the delayers 604-1, 604-2, the replica signals of the communication partner 1 are eliminated from the received signals delayed by the delayers 604-1, and' 604-2.

[0068] In the ICU 606-2, a reception weight, a steering vector, and a replica signal of the communication partner 2 are generated every antenna based on a signal obtained by eliminating the replica signal of the communication partner 1, from the received signal. The replica

signal of the communication partner 2 generated every antenna is inputted to each of next adders 607-1 and 607-2, and the steering vector every antenna is inputted to the ICU 606-2 of the second stage.

[0069] In the adders 607-1 and 607-2 connected to the delayers 605-1, 605-2, the replica signals of the communication partner 1 and those of the communication partner 2 are eliminated from the received signals delayed by the delayers 605-1, and 605-2.

[0070] Then, in the ICU 606-3, a reception weight, a steering vector, and a replica signal of the communication partner 3 are generated every antenna based on a signal obtained by eliminating the replica signal of the communication partner 1 and the replica signal of the communication partner 2 from the received signal.

[0071] In this way, since each ICU in one stage calculates the reception weight based on the signal from which the interference signals are sequentially eliminated, the reception weights are sequentially updated every communication partner in one stage.

[0072] An explanation will be next given of the radiation pattern generated by each ICU of the first stage using FIGS. 7A to 7C. FIG. 7A to FIG. 7C are view each showing an example of a radiation pattern formed by each ICU of the radio receiving apparatus according to Embodiment 2 of the present invention. In FIGS. 7A to 7C, it is assumed that the wider the width of the arrow becomes, the larger transmission power becomes.

[0073] First, all signals sent from the communication partners 1 to 3 are contained in the signals inputted in the ICU 606-1. It is assumed that the ICU 606-1 performs the array reception using the null steering. In the ICU 606-1, as shown in FIG. 7A, the radiation pattern is generated in such a way that the null point is directed to the direction where the communication partner 2 exists. This makes it possible for the ICU 606-1 to generate the replica signal after eliminating interference received from the communication partner 2. As a result, the replica signal of the communication partner 1 can be accurately generated.

[0074] The reason why the null point is not directed to the direction where the communication partner 3 exists is as follows:

[0075] Since the number of antennas is two, the number of null points that can be generated is only one, with the result that the null point is formed in the direction where the communication partner 2 providing a large quality of interference exists.

[0076] Since the replica signal of the communication partner 1 is eliminated from the received signal by the adders 607-1 and 607-2 connected to the delayers 604-1 and 604-2, only the signals sent from the communication partners 2 and 3 are contained in the signals inputted to the ICU 606-2. Accordingly, in the ICU 606-2, as illustrated in FIG. 7B, the radiation pattern is generated in such a way that the null point is directed to the direction where the communication partner 3 exists. This makes it possible for the ICU 606-2 to generate the replica signal after eliminating interference received from the communication partner 3 from the signals from which interference received from the communication partner 1 is eliminated. As a result, the replica signal of the communication partner 2 can be accurately generated.

[0077] Then, since the replica signals of the communication partners 1 and 2 are eliminated from the received signals by the adders 607-1 and 607-2 connected to the delayers 605-1 and 605-2, only the signal sent from the communication partner 3 is inputted to the ICU 606-3. Accordingly, in the ICU 606-3, as illustrated in FIG. 7C, the radiation pattern is generated in such a way that the beam point is directed to the direction where the communication partner 3 exists. This makes it possible to generate the replica signal of the communication partner 3 accurately.

[0078] Embodiment 1 has explained the radio receiving apparatus in which the array antenna and the parallel type interference canceller are combined. In the radio receiving apparatus of embodiment 1, the parallel type interference canceller is used, so that the inference signals of the respective communication partners are simultaneously eliminated in parallel in one stage. For this reason, in Embodiment 1, each ICU in one stage calculates the reception weights without considering the interference signals to be eliminated in the stage.

[0079] In contrast to this, the radio receiving apparatus of Embodiment 2 is the radio receiving apparatus in which the array antenna and the serial type interference canceller are combined as illustrated in FIG. 6. For this reason, in the radio receiving apparatus of this embodiment, the interference signals are sequentially eliminated for every communication partner in one stage. Accordingly, in the radio receiving apparatus of this embodiment, the signals from which interference signals are sequentially eliminated are inputted to each ICU in one stage.

[0080] In other words, each ICU of the radio receiving apparatus of this embodiment calculates the reception weights with respect to the signals from which the interference signals are sequentially eliminated in one stage. Hence, as compared with each ICU of the radio receiving apparatus of Embodiment 1, it is possible to calculate the reception weights with respect to the signals having a small amount of interference. Accordingly, the radio receiving apparatus of this embodiment can generate the radiation pattern and the replica signal more accurately as compared with Embodiment 1. This makes it possible to obtain high interference cancellation capability even if the number of stages is further reduced as compared with Embodiment 1. Therefore, it is possible to further reduce the apparatus scale.

[0081] Thus, according to the radio receiving apparatus and the radio receiving method according to this embodiment, the interference signals are sequentially eliminated every communication partner in one stage to update the reception weights sequentially every commu-

nication partner in one stage. This makes it possible to improve the accuracy of the radiation pattern and that of the replica signal. Therefore, according to the radio receiving apparatus and the radio receiving method according to this embodiment, it is possible to obtain high interference cancellation capability even if the number of stages is further reduced as compared with Embodiment 1, and this makes it possible to further reduce the apparatus scale.

[0082] Additionally, in Embodiments 1 and 2, the method for estimating the direction of arrival is not particularly limited. The estimation of the direction of arrival aims to obtain the steering vector S_k every antenna. For this reason, the radio receiving apparatus of Embodiment 1 and 2 may obtain the steering vector S_k using any method as long as the steering vector S_k can be obtained. For example, the radio receiving apparatus of Embodiment 1 and 2 calculate the correlation value between the signal received by each antenna and the known signal to make it possible to obtain the steering vector S_k.

[0083] Embodiment 1 has explained the radio receiving apparatus in which the array antenna and the parallel type interference canceller are combined. Embodiment 2 has explained the radio receiving apparatus in which the array antenna and the serial type interference canceller are combined. However, the present invention can be applied to the radio receiving apparatus in which the array antenna and the symbol ranking type interference canceller are combined.

[0084] As explained above, according to the present invention, even if the array antenna and the interference canceller are combined, it is possible to receive a desired signal with high quality in an apparatus of small scale without providing an interference canceller to each channel corresponding to each communication partner.

[0085] This application is based on the Japanese Patent Application No. 2000-010878 filed on January 19, 2000, entire content of which is expressly incorporated by reference herein

Industrial Applicability

[0086] The present invention is suitable for use in a mobile station apparatus and a base station apparatus in a mobile communication system. In the case of application, it is possible to receive a desired signal with high quality in an apparatus of' small scale even if the array antenna and the interference canceller are combined in the mobile station apparatus and the base station apparatus.

Claims

1. A radio receiving apparatus comprising:

    a first calculation section for calculating reception weighting factors with respect to received signals received by the respective antenna element composing an adaptive array antenna;
    an arrival direction estimation section for estimating directions of arrival of said received signals;
    a second calculation section for calculating weighting factors for a replica signal generation in accordance with said reception weighting factors and said directions of arrival;
    a replica signal generator for generating replica signals of each of said received signals using said weighting factors for a replica signal generation; and
    an eliminator for eliminating said replica signals from said received signals.

2. The radio receiving apparatus according to claim 1, wherein

    said first calculation section calculates reception weighting factors by which a radiation pattern is formed in such a way that a null point is directed to a direction where an interference signal source exists.

3. The radio receiving apparatus according to claim 1, comprising

    a plurality of processors each having said first calculation section, said arrival direction estimation section, and said eliminator, as a multistage.

4. The radio receiving apparatus according to claim 3, wherein

    in the processor of a latter stage,
    said first calculation section calculates the reception weighting factors with respect to the signals obtained by eliminating the replica signals from the received signals by said eliminator in a preceding stage, whereby updating the reception weighting factors sequentially.

5. The radio receiving apparatus according to claim 3, wherein

    in the processor of a latter stage,
    said arrival direction estimation section estimates the directions of arrival of the signals obtained by eliminating the replica signals from the received signals by said eliminator in a preceding stage.

6. The radio receiving apparatus according to claim 5, wherein

in the processor of a latter stage,
said arrival direction estimation section estimates the directions of arrival using an average value of calculated steering vectors in a given interval.

7. A mobile station apparatus having a radio receiving apparatus thereon, said radio receiving apparatus comprising:

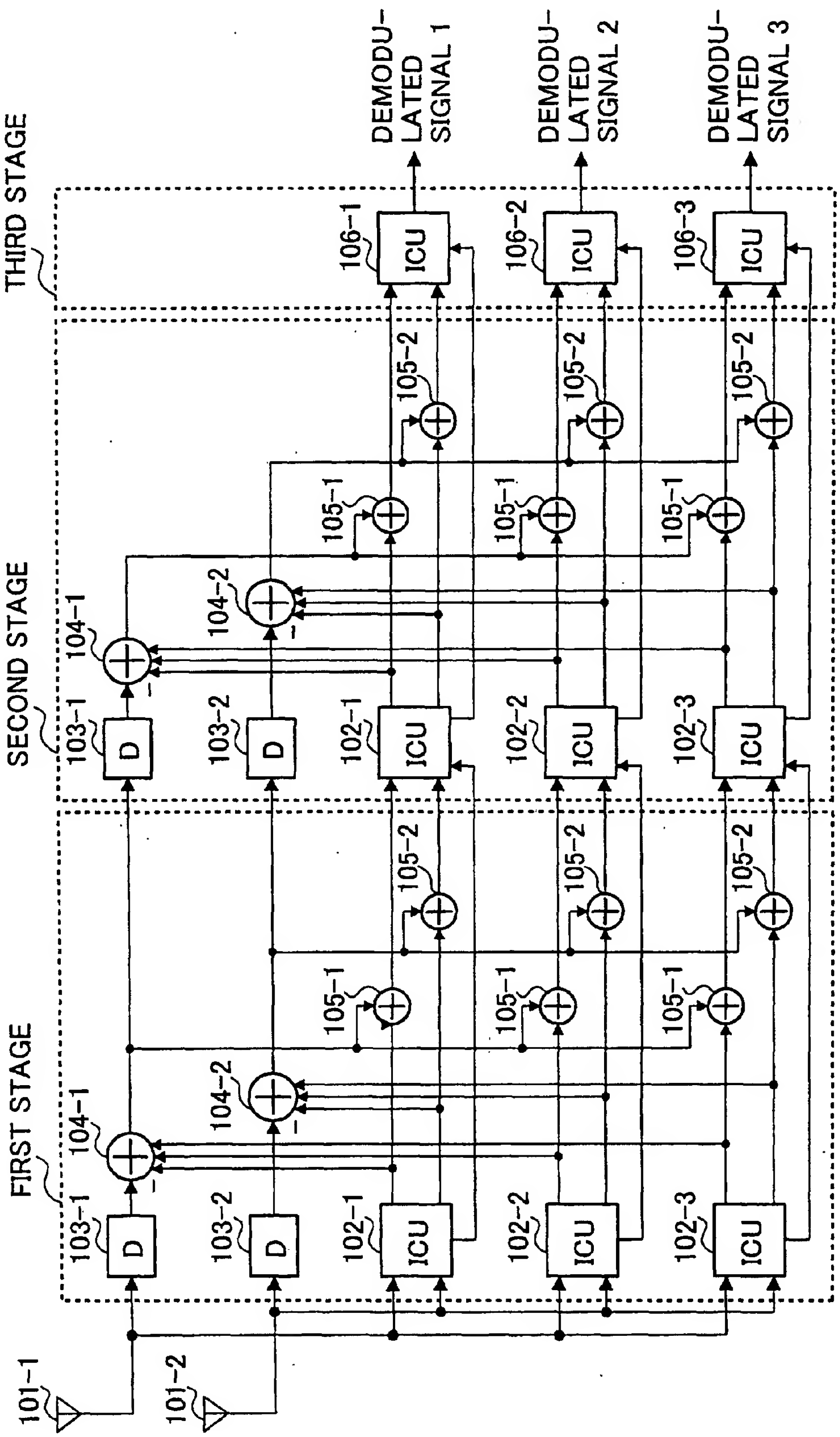
a first calculation section for calculating reception weighting factors with respect to received signals received by the respective antenna element composing an adaptive array antenna;
an arrival direction estimation section for estimating directions of arrival of said received signals;
a second calculation section for calculating weighting factors for a replica signal generation in accordance with said reception weighting factors and said directions of arrival;
a replica signal generator for generating replica signals of each of said received signals using said weighting factors for a replica signal generation; and
an eliminator for eliminating said replica signals from said received signals.

8. A base station apparatus having a radio receiving apparatus thereon, said radio receiving apparatus comprising:

a first calculation section for calculating reception weighting factors with respect to received signals received by the respective antenna element composing an adaptive array antenna;
an arrival direction estimation section for estimating directions of arrival of said received signals;
a second calculation section for calculating weighting factors for a replica signal generation in accordance with said reception weighting factors and said directions of arrival;
a replica signal generator for generating replica signals of each of said received signals using said weighting factors for a replica signal generation; and
an eliminator for eliminating said replica signals from said received signals.

9. A radio receiving method comprising the steps of:

calculating reception weighting factors with respect to received signals received by the respective antenna element composing an adaptive array antenna;
estimating directions of arrival of said received signals;
calculating weighting factors for a replica signal generation in accordance with said reception weighting factors and said directions of arrival;
generating replica signals of each of said received signals using said weighting factors for a replica signal generation; and
eliminating said replica signals from said received signals.



FIRST STAGE
SECOND STAGE
THIRD STAGE
101-1
101-2
103-1
D
103-2
D
104-1
104-2
102-1
ICU
102-2
ICU
102-3
ICU
105-1
105-2
106-1
ICU
106-2
ICU
106-3
ICU
DEMODU-
LATED
SIGNAL 1
DEMODU-
LATED
SIGNAL 2
DEMODU-
LATED
SIGNAL 3

FIG.1

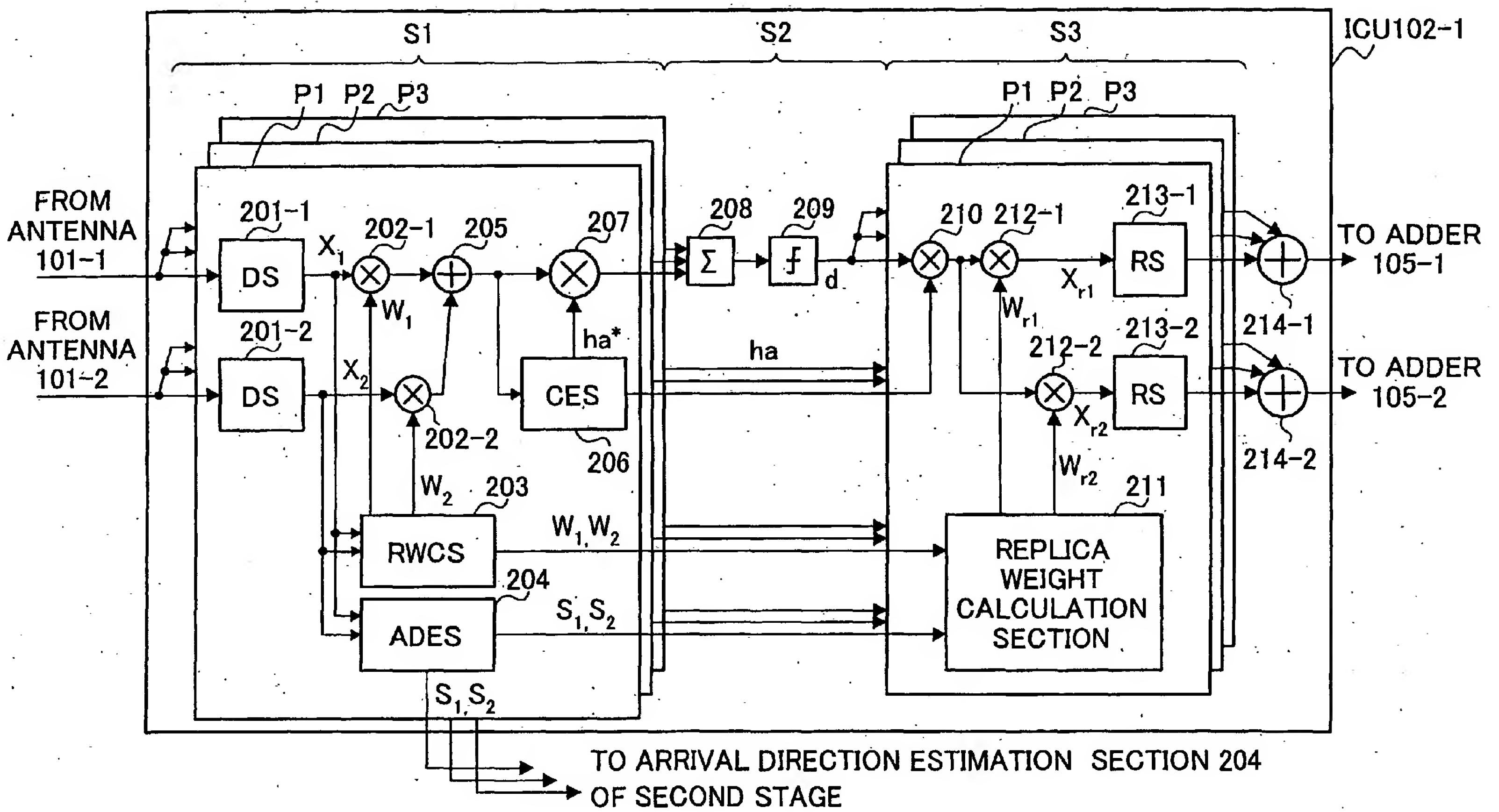


FIG.2

201 DS : DESPREADING SECTION
203 RWCS : RECEPTION WEIGHT CALCULATION SECTION
204 ADES : ARRIVAL DIRECTION ESTIMATION SECTION
206 CES : CHANNEL ESTIMATION SECTION
213 RS : RE-SPREADING SECTION

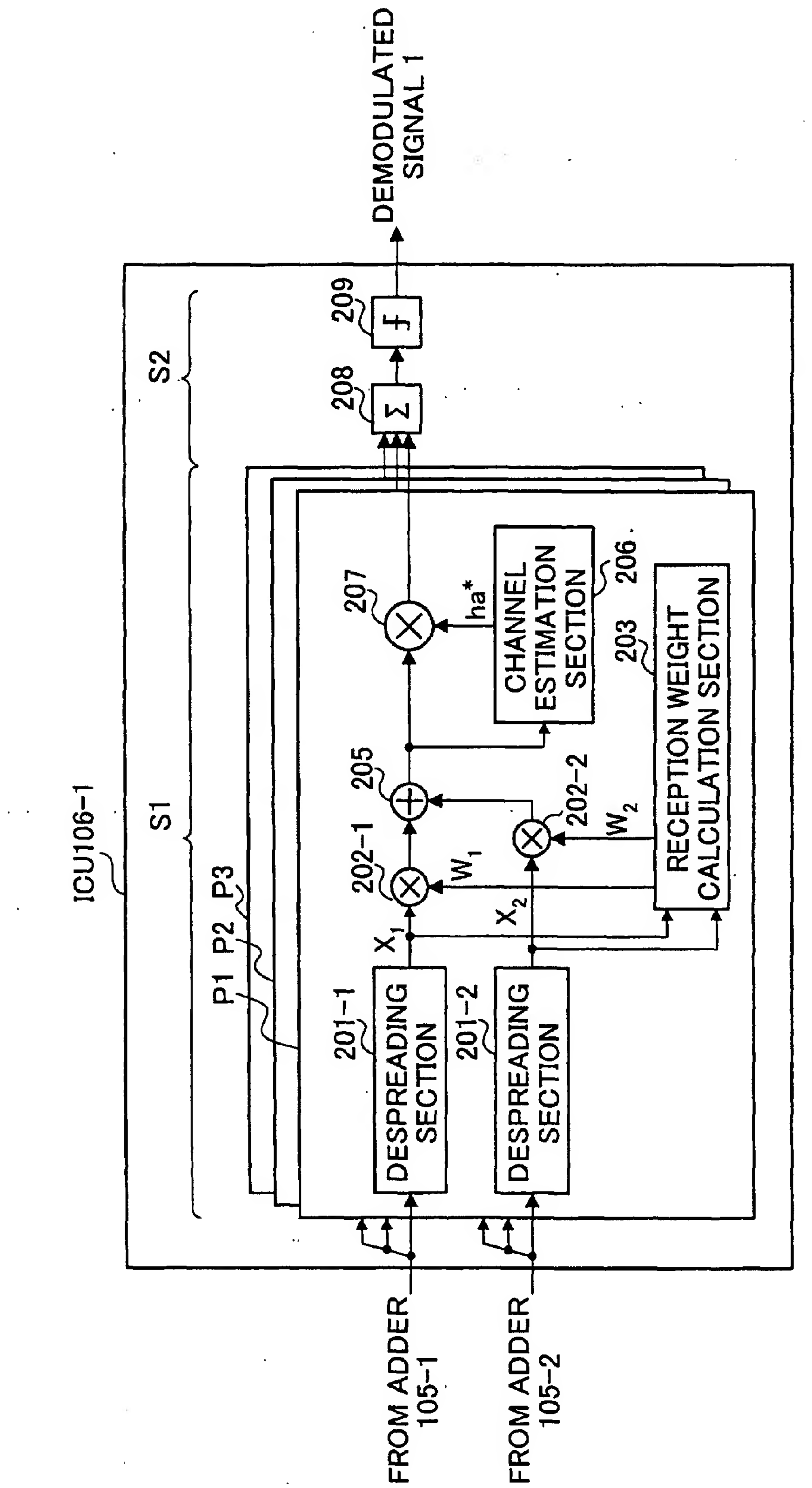

ICU106-1
S1
S2
P1 P2 P3
FROM ADDER 105-1
FROM ADDER 105-2
201-1
DESPREADING SECTION
201-2
DESPREADING SECTION
X_1
X_2
202-1
202-2
W_1
W_2
203
RECEPTION WEIGHT CALCULATION SECTION
205
206
CHANNEL ESTIMATION SECTION
ha*
207
208
Σ
209
DEMODULATED SIGNAL 1

FIG.3

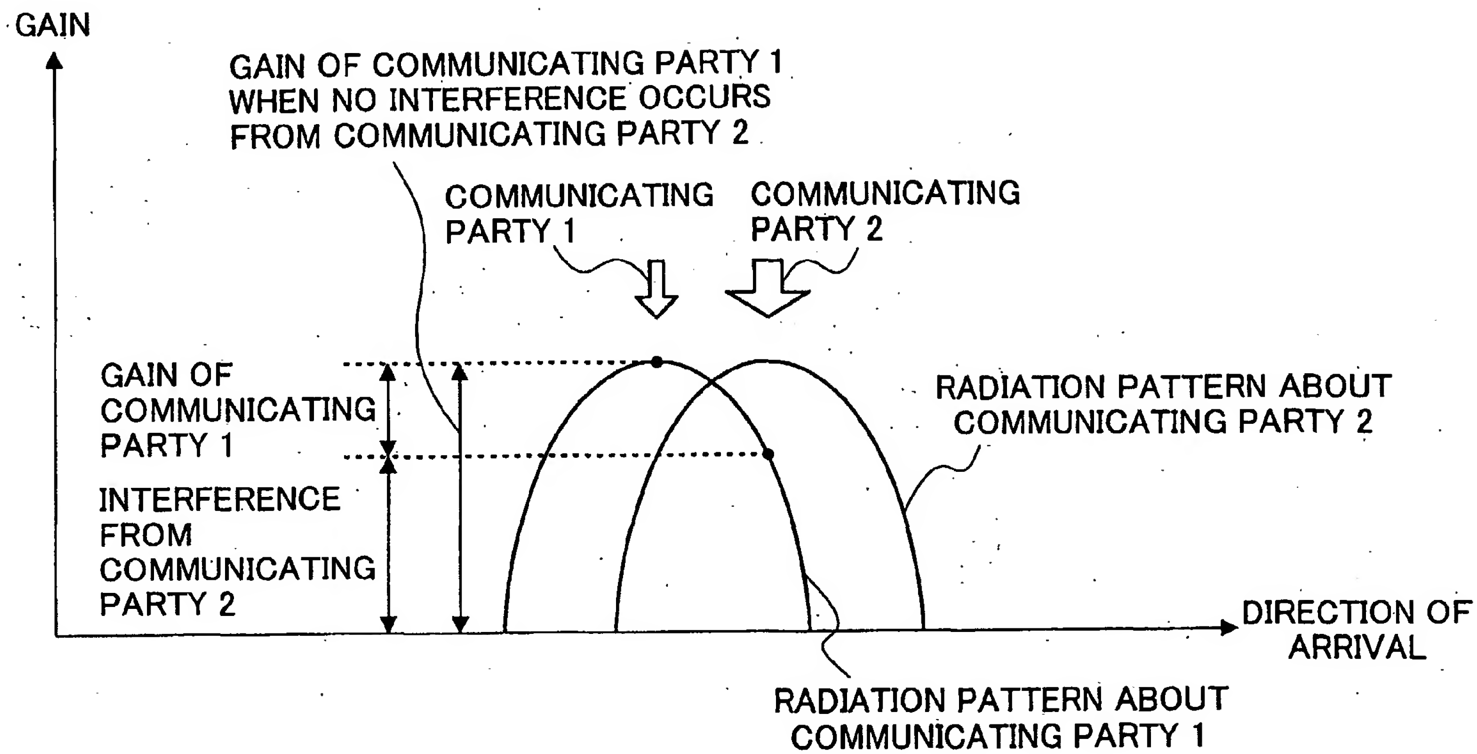

GAIN
GAIN OF COMMUNICATING PARTY 1
WHEN NO INTERFERENCE OCCURS
FROM COMMUNICATING PARTY 2
COMMUNICATING PARTY 1
COMMUNICATING PARTY 2
GAIN OF COMMUNICATING PARTY 1
INTERFERENCE FROM COMMUNICATING PARTY 2
RADIATION PATTERN ABOUT COMMUNICATING PARTY 2
DIRECTION OF ARRIVAL
RADIATION PATTERN ABOUT COMMUNICATING PARTY 1

FIG.4

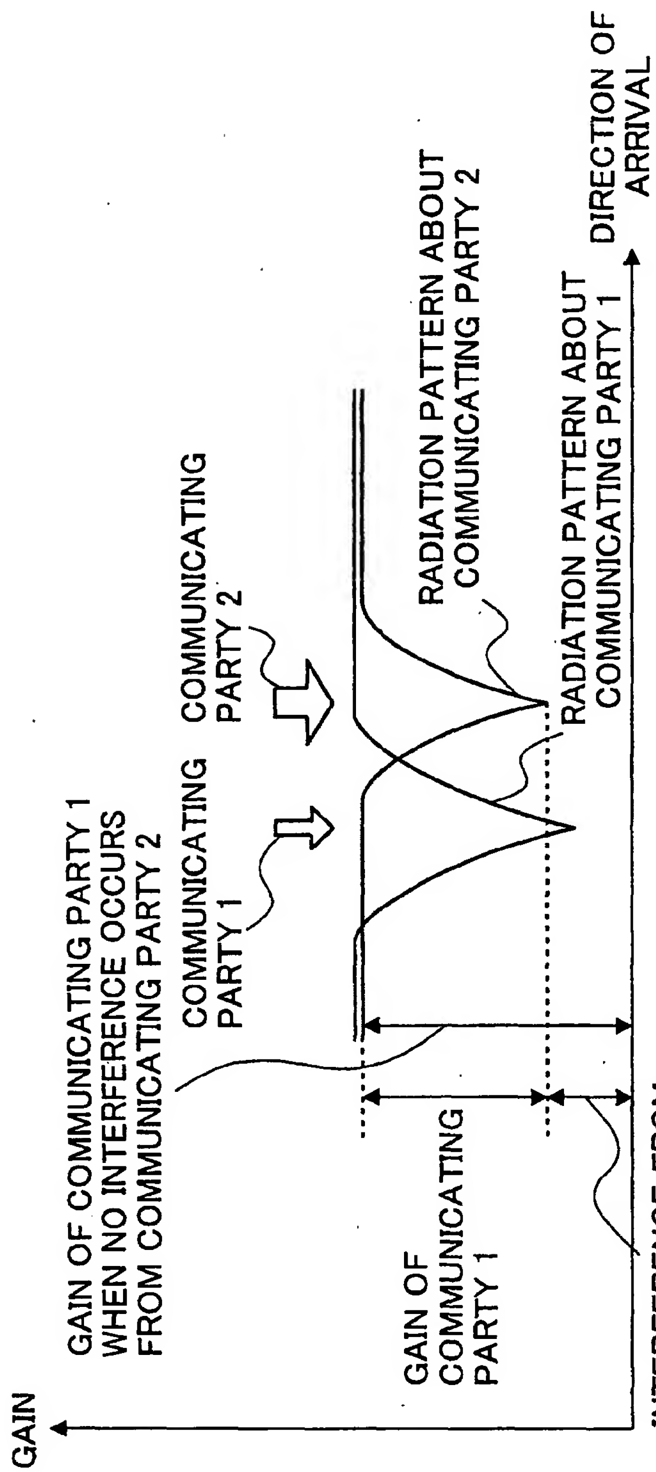
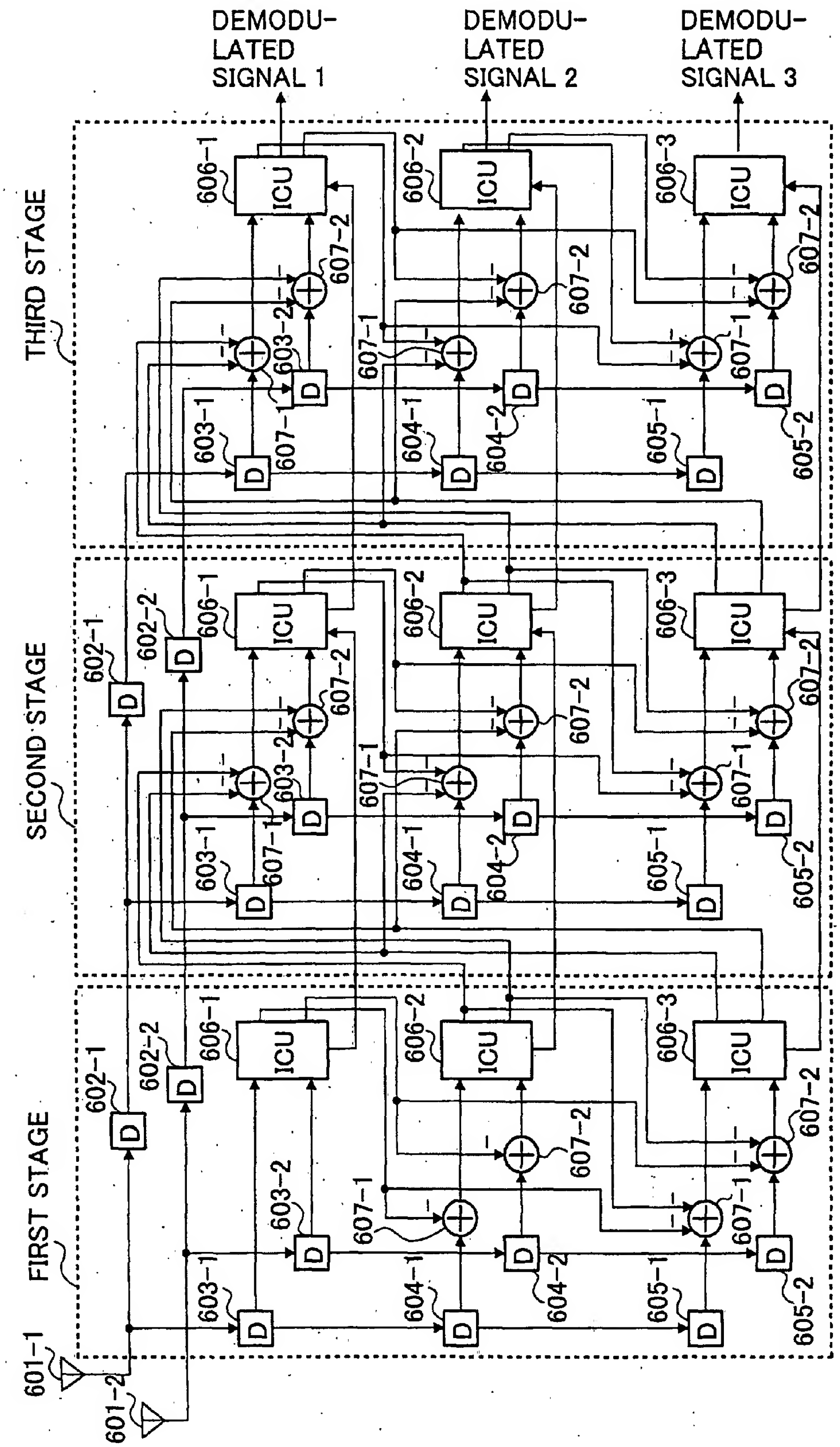

GAIN
GAIN OF COMMUNICATING PARTY 1 WHEN NO INTERFERENCE OCCURS FROM COMMUNICATING PARTY 2
COMMUNICATING PARTY 1
COMMUNICATING PARTY 2
GAIN OF COMMUNICATING PARTY 1
RADIATION PATTERN ABOUT COMMUNICATING PARTY 2
RADIATION PATTERN ABOUT COMMUNICATING PARTY 1
INTERFERENCE FROM COMMUNICATING PARTY 2
DIRECTION OF ARRIVAL

FIG.5


DEMODU-
LATED
SIGNAL 1
DEMODU-
LATED
SIGNAL 2
DEMODU-
LATED
SIGNAL 3
THIRD STAGE
SECOND STAGE
FIRST STAGE
ICU
D
601-1
601-2
602-1
602-2
603-1
603-2
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604-2
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606-3
607-1
607-2

FIG.6

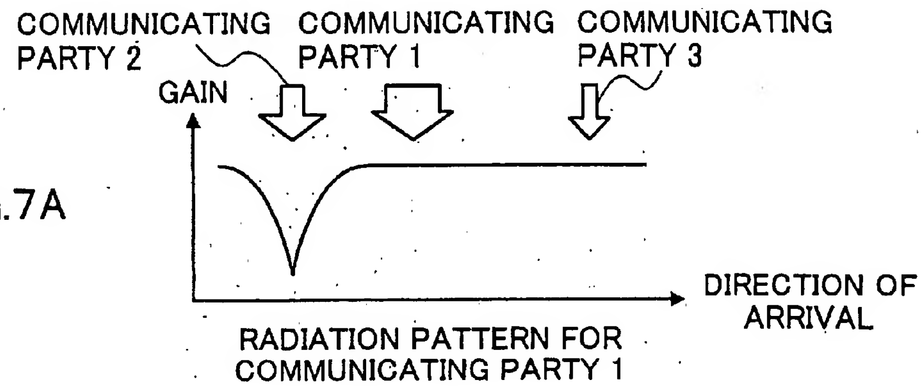

FIG.7A

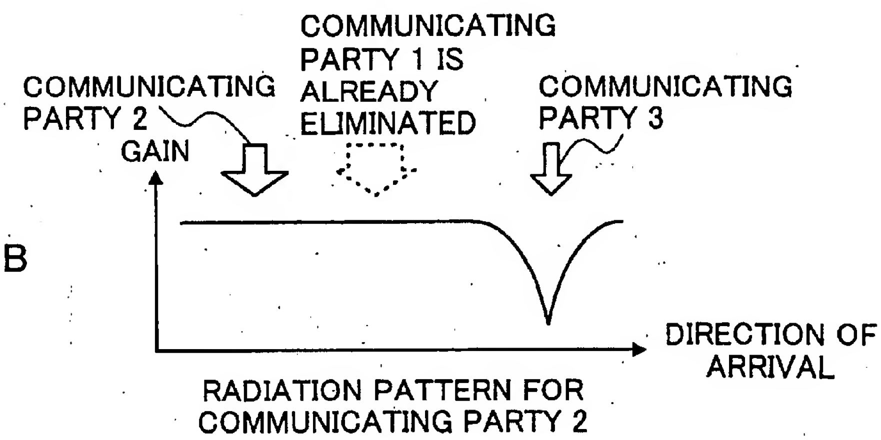

FIG.7B

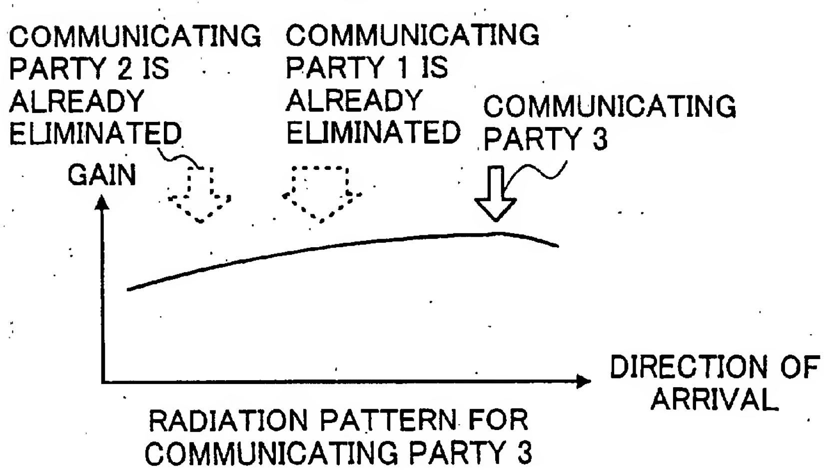

FIG.7C

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP01/00250

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl^7 H04B7/08, 7/10, 1/10, 7/26
H01Q3/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl^7 H01Q3/00-3/46, 21/00-25/04
H04B7/00, 7/02-7/12, 7/24-7/26, H04J13/04
H04Q7/00-7/38, H04L1/02-1/06

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
Jitsuyo Shinan Koho 1922-1996 Toroku Jitsuyo Shinan Koho 1994-2001
Kokai Jitsuyo Shinan Koho 1971-2001 Jitsuyo Shinan Toroku Koho 1996-2001

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
E,X E,A	JP, 2001-16148, A (Fujitsu Limited), 19 January, 2001 (19.01.01) (Family: none)	1-5,7-9 6
X A	JP, 11-205286, A (NEC Corporation), 30 July, 1999 (30.07.99) (Family: none)	1,3-5,7-9 2,6
P,A	JP, 2000-138605, A (NEC Corporation), 16 May, 2000 (16.05.00) & WO, 200027062, A1	1-9
A	JP, 11-251959, A (Fujitsu Limited), 17 September, 1999 (17.09.99) (Family: none)	1-9
A	JP, 10-190495, A (Fujitsu Limited), 21 July, 1998 (21.07.98) & EP, 849888, A & US, 6157685, A	1-9

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
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"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
22 March, 2001 (22.03.01)	03 April, 2001 (03.04.01)
Name and mailing address of the ISA/ Japanese Patent Office	Authorized officer
Facsimile No.	Telephone No.

Form PCT/ISA/210 (second sheet) (July 1992)

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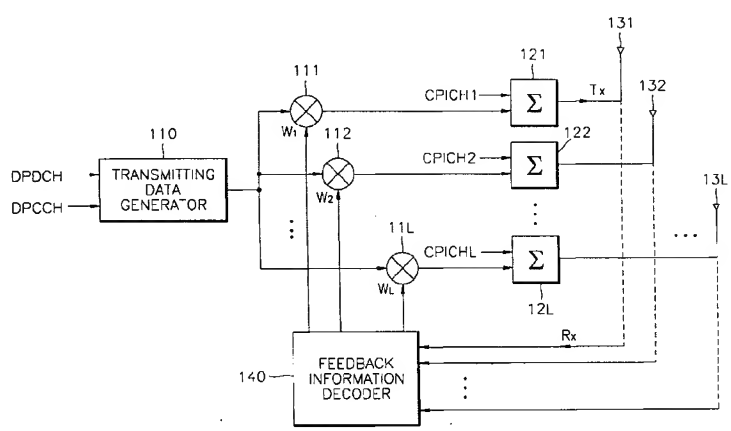
(54) Transmission antenna diversity method, base station and mobile station therefor

(57) A transmission antenna diversity method, and a base station apparatus and a mobile station apparatus therefor in a mobile communication system are provided. In the transmission antenna diversity method, channel information is measured from signals received through the plurality of antennas used in the base station, and a channel information matrix is output. The channel information matrix is transformed according to a transform matrix composed of a complex basis vector set. Reception power with respect to the plurality of antennas is calculated based on the transformed channel information matrix. Antenna selection information obtained based on the calculated reception power is transmitted to the base station as feedback information for controlling transmission antenna diversity. Therefore, power is equally distributed to transmitting antennas, excellent performance is maintained at a high speed of movement, and reliable channel adaptation is accomplished at a low speed of movement.

FIG. 1



EP 1 175 022 A2

Description

[0001] The present invention relates to transmission antenna diversity, and more particularly, to a transmission antenna diversity method, and a base station apparatus and a mobile station apparatus therefor in a mobile communication system.

[0002] Third generation mobile communication systems have standards for transmitting data at a higher rate than second generation mobile communication systems represented by personal communication systems (PCS). In Europe and Japan, a synchronous Wideband Code Division Multiple Access (W-CDMA) mode is adopted as a wireless access standard. In North America, an asynchronous CDMA-2000 mode is adopted as a wireless access standard. Mobile communication systems are configured so that many mobile stations can communicate through a single base station.

[0003] It is necessary to overcome fading in order to transmit data at a high rate in a mobile communication system. Fading reduces the amplitude of a received signal by several decibels to several tens of decibels. A variety of diversity techniques are used to satisfactorily overcome fading. In a CDMA mode, a Rake receiver using delay spread of a channel is employed. A reception diversity technique of receiving a multi-path signal is applied to a Rake receiver. This reception diversity technique has a problem in that reception diversity does not operate well when delay spread is small. A time diversity technique using interleaving and coding is used in a Doppler spread channel. It is difficult to use this method in a low-speed Doppler channel.

[0004] Space diversity is used to overcome fading in an indoor channel having small delay spread and an outdoor channel which is a low-speed Doppler channel. Space diversity uses two or more antennas. In this method, when a signal transmitted through one antenna is reduced due to fading, a signal transmitted through another antenna is used for reception. Space diversity is divided into reception antenna diversity using receiving antennas and transmission antenna diversity using transmitting antennas. It is difficult to install reception antenna diversity in a mobile station because of lack of space and excessive cost, so it is recommended to use transmission antenna diversity in a base station.

[0005] Transmission antenna diversity is divided into closed loop transmission diversity of feeding back uplink channel information from a mobile station for operation and open loop transmission diversity not having feedback from a mobile station. When using L antennas, closed loop transmission diversity has L times greater gain than open loop transmission diversity in terms of a Signal to Interference and Noise Ratio (SINR). However, the performance of closed loop transmission diversity of feeding back channel information for operation is influenced by the period of feedback. When the period of feedback is long, a channel may change before feedback information reaches a base station, thereby decreasing the performance. When a large amount of information is fed back per unit time in order to follow up a rapidly changing channel, up-link capacity decreases.

[0006] In addition, transmission antenna diversity is divided into a maximal ratio combining (MRC) method, an equal gain combining (EGC) method, and a selective combining (SC) method. When a feedback bandwidth is not satisfactorily secured, the performance of the above-described closed loop transmission antenna diversity may be deteriorated because a change in channel information is not reliably reflected in feedback information. Here, in order to make channel information to be rapidly and reliably reflected in feedback information rather than to obtain exact channel information, closed loop transmission antenna diversity employing an SC method is used.

[0007] However, when using an SC method, unbalance between antennas occurs. Accordingly, it costs a lot to configure a radio frequency (RF) processor. Therefore, diversity using an SC method for overcoming the above problem and performing diversity using less feedback information is desired.

[0008] Although diversity gain can be obtained, SINR gain decreases in diversity using an SC method compared to diversity using an MRC method or an EGC method because channel information is not completely reflected by feedback information. Therefore, an improved diversity method which can maximize SINR gain by compensating for the decrease, can be applied at a high moving speed, and can simplify the hardware configuration of a transceiver is desired.

[0009] A transmission antenna diversity method using a feedback mode is disclosed in U.S. Patent Nos. 5,634,199 and 5,471,647. In these patents, measurement of a channel and a feedback method using a perturbation algorithm and a gain matrix are proposed. However, these patents adopt a blind method, which is slow in convergence for measuring a channel and is not effective in finding an exact weight, so the method is not frequently used in systems using a pilot signal.

[0010] In Universal Mobile Telecommunication Service (UMTS) W-CDMA (3GPP) standards, Motorola proposes a method of quantizing a weight for each antenna in a feedback mode. In addition, Nokia and others propose a transmission antenna diversity method for high-speed mobile objects operating with respect to two antennas. However, these methods are optimized for a case of using two antennas. Therefore, an improved method of effectively selecting many antennas is desired.

[0011] According to a first aspect of the invention there is provided a closed loop transmission antenna diversity method employing a selective combining method, including the steps of (a) measuring channel information from signals received through a plurality of antennas used in a base station and outputting a channel information matrix, (b) transforming the channel information matrix according to a transform matrix com-

posed of a complex basis vector set, (c) calculating reception power with respect to the plurality of antennas based on the transformed channel information matrix, and (d) transmitting antenna selection information obtained based on the calculated reception power to the base station as feedback information for controlling transmission antenna diversity.

[0012] In another aspect, there is provided a closed loop transmission antenna diversity method employing a selective combining method, including the steps of (a) receiving selection information related to a complex basis vector from a mobile station as feedback information for controlling transmission antenna diversity; (b) determining a complex basis vector selected based on the selection information; (c) obtaining an antenna weight for each antenna using the determined complex basis vector; and (d) generating a signal based on the antenna weight and transmitting the signal to the mobile station through a corresponding antenna.

[0013] The present invention accordingly provides a closed loop transmission antenna diversity method using a selective combining (SC) method, which can overcome an imbalance of power between antennas by providing feedback information for selecting an antenna using a complex basis vector set in extending antenna selection to basis vector selection.

[0014] In another aspect of the invention there is provided a base station apparatus including a plurality of antennas for receiving selection information related to a complex basis vector from a mobile station as feedback information for controlling transmission antenna diversity, a feedback information decoder for determining a complex basis vector selected based on the selection information and obtaining an antenna weight for each antenna using the determined complex basis vector, and a data transmitting unit for generating a signal based on the antenna weight and transmitting the signal to the mobile station through a corresponding antenna.

[0015] In a further aspect of the invention there is provided a mobile station apparatus including a channel information measuring unit for measuring channel information from signals received through a plurality of antennas used in a base station and outputting a channel information matrix, a basis vector transformer for transforming the channel information matrix according to a transform matrix composed of a complex basis vector set, an optimum weight detector for calculating reception power with respect to the plurality of antennas based on the transformed channel information matrix and generating feedback information for controlling transmission antenna diversity based on the calculated reception power, and an uplink signal processor for transmitting the feedback information to the base station in the form of a symbol configured according to a protocol suitable for feedback.

[0016] The present invention accordingly provides a base station apparatus and a mobile station apparatus for performing the above closed loop transmission antenna diversity method using an SC method.

[0017] The objects and advantages of the present invention will become more apparent by describing in detail preferred embodiments thereof with reference to the attached drawings in which:

FIG. 1 is a block diagram of a transmitting apparatus for transmission antenna diversity in a wireless communication system;

FIG. 2 is a block diagram of an embodiment of the feedback information decoder shown in FIG. 1;

FIG. 3 is a block diagram of a receiving apparatus for transmission antenna diversity in a wireless communication system;

FIG. 4 is a block diagram of an embodiment of a receiving apparatus for transmission antenna diversity in a wireless communication system;

FIGS. 5A through 5C show examples of a basis vector set on a real axis, a basis vector set on an imaginary axis, and a complex basis vector set obtained by combining the two real and imaginary basis vector sets;

FIG. 6 shows the mapping relation between a basis vector and feedback information at each slot number; and

FIG. 7 shows parameters and their values used for transmission antenna diversity employing a selection method using complex basis vectors when four antennas are used.

[0018] The operating principle of the present invention will now be briefly described. The present invention employs a selective combining (SC) method for closed loop transmission antenna diversity in a wireless transmit-receive system. In the case of transmission antenna diversity in which a transmitter transmits a signal through a plurality of antennas, the SC method simplifies hardware configuration. However, due to the imbalance of power between antennas, cost for configuring a radio frequency (RF) processor increases. To overcome this problem, selection of an antenna is extended to selection of a basis vector. In selection of a basis vector, by using an equal power balanced basis vector allowing antennas to have the same power, the imbalance of power between antennas can be overcome even if diversity employing an SC method is performed.

[0019] A basis vector set, { [1 0 0 0], [0 1 0 0], [0 0 1 0], [0 0 0 1]}, may be used to select an antenna in a receiver. This set is an unequal power balanced basis vector set. Equal power balanced basis vector sets are a Walsh basis vector set, { [1 1 1 1], [1 -1 1 -1], [1 1 -1 -1], [1 -1 -1 1]}, and a polar basis vector set, { [-1 1 1 1], [1 -1 1 1], [1 1 -1 1], [1 1 1 -1]}.

[0020] These basis vector sets used for an SC method are composed of the same constants so that the inner product of different vectors is 0, and the inner product of the same vectors is not 0. When these basis vector sets are used to obtain weights for antennas, vectors

are equalized so that the constant can be 1 in order not to change transmission power. Such an equalized set is referred to as an orthonormal basis vector set.

[0021] For reference, since a method of using an equal power balanced orthonormal basis vector set in reception antenna diversity comes under diversity based on an SC method, the performance of the method is the same as that of a method using an unequal power balanced orthonormal basis vector set. When it is assumed that diversity information is ideally fed back, both methods have the same performance in transmission antenna diversity, with the exception that power is uniform among transmitting antennas in the method of using an equal power balanced orthonormal basis vector set.

[0022] In closed loop transmission antenna diversity configured so that diversity information is fed back from a mobile station, transmission of feedback information is delayed as the moving speed of a mobile station increases because the bandwidth of a feedback channel is limited. This delay decreases diversity gain. As compared to an SC method, a maximal ratio combining (MRC) method or an equal gain combining (EGC) method has a large amount of feedback information so that accurate channel compensation can be performed at a low moving speed, thereby increasing the performance, but the performance rapidly decreases as the moving speed of the mobile station increases.

[0023] The present invention provides a complex basis vector selection method for improving the performance of diversity using an SC method at a low moving speed and maintaining the performance as the moving speed increases. A complex basis vector set is composed of different orthonormal sets assigned to the real axis and the imaginary axis of the complex plane. For example, a Walsh basis vector set is assigned to a real axis, and a polar basis vector set is assigned to an imaginary axis. When four antennas are used, a complex basis vector set composed of 16 vector combinations is obtained. Selecting one complex basis vector from the complex basis vector set is determining an antenna to be given a weight among a plurality of antennas.

[0024] When information related to selection of a complex basis vector is fed back to a base station from a mobile station, vector information as to a real axis and vector information as to an imaginary axis is alternately transmitted at feedback signalling intervals. The base station sums information received during two consecutive feedback signalling intervals by way of sliding window and forms a complex basis vector. For example, when feedback information is transmitted in order of real axis information and imaginary axis information, one basis vector is constituted by the first real axis information and the second imaginary axis information, and then the next basis vector is constituted by the second imaginary axis information and the third real axis information, thereby forming a complex basis vector by way of sliding window. Each factor of the complex basis vector is used as a weight for each antenna. By configuring transmission antenna diversity as described above, optimum feedback information can be used at feedback signalling intervals. Since an optimum weight can be used at feedback signalling intervals, excellent characteristics are maintained at a high moving speed, and the performance is improved at a low moving speed as the resolution of the complex basis vector is increased to 1/16. Furthermore, the complex basis vector set is formed to equalize power among antennas, thereby preventing an imbalance of power among the antennas.

[0025] Hereinafter, the configurations and operations of a transmitting apparatus and a receiving apparatus according to the present invention will be specifically described. FIG. 1 is a block diagram of a transmitting apparatus for transmission antenna diversity in a wireless communication system. The transmitting apparatus corresponds to a base station in a mobile communication system and can be referred to as a UMTS (Universal Mobile Telecommunication Service) Terrestrial Radio Access Network (UTRAN).

[0026] Referring to FIG. 1, the transmitting apparatus includes a transmitting data generator 100, L (two or more) multipliers 111 through 11L, L adders 121 through 12L, L antennas 131 through 13L, and a feedback information decoder 140. The transmitting data generator 100 generates and outputs data to be transmitted to the L multipliers 111 through 11L. Specifically, the transmitting data generator 100 receives, for example, a signal DPDCH of a dedicated physical data channel and a signal DPCCH of a dedicated physical control channel, and multiplexes these signals to generate and output transmitting data.

[0027] The L multipliers 111 through 11 L multiply the data output from the transmitting data generator 100 by weights w_1 through w_L corresponding to the respective antennas 131 through 13L. The L adders 121 through 12L add the pilot signals CPICH1 through CPICHL corresponding to the respective antennas 131 through 131 L to the outputs of the corresponding L multipliers 111 through 11L, respectively. The signals generated by the L adders 121 through 12L are forwarded through the corresponding L antennas 131 through 13L, respectively, via a radio frequency (RF) signal processor (not shown).

[0028] Here, the antenna weights w_1 through w_L are obtained through the operation of the feedback information decoder 140 analyzing feedback information received through the L antennas 131 through 13L. Feedback information is uplinked from a receiving apparatus (that is, an arbitrary i-th mobile station). Practically, the L antennas 131 through 13L receive as feedback information an index indicating one element (i.e., one complex basis vector) of a complex basis vector set. This will be described in detail later.

[0029] The feedback information decoder 140 selects a complex basis vector corresponding to the index received as feedback information. Factors of the selected

complex basis vector are output as weights corresponding to the respective L antennas 131 through 13L.

[0030] FIG. 2 is a block diagram of an embodiment of the feedback information decoder 140 shown in FIG. 1. The feedback information decoder 140 includes a switching unit 200, a feedback signaling message (FSM) register 210, first and second weight tables 222 and 224, and an adder 230.

[0031] The switching unit 200 operates according to whether the slot number of a received signal is even or odd so that feedback information can be stored in a real number portion of the FSM register 210 when the slot number is even and in an imaginary number portion of the FSM register 210 when the slot number is odd. Here, the feedback information is an index indicating one basis vector in a first orthonormal basis vector set indicating a real part or an index indicating one basis vector in a second orthonormal basis vector set indicating an imaginary part. When each basis vector set is composed of four basis vectors, a feedback signaling vector $[m_{b,1}, m_{b,2}]^T$ transmitted from a mobile station to a base station is represented by 2-bit binary data expressing an index indicating a basis vector.

[0032] The FSM register 210 outputs an index i_w, which will be input to the first weight table (i.e., a look-up table) 222, using feedback information (e.g., an index represented by two bits) stored in the real number portion, and an index i_p, which will be input to the second weight table 224, using feedback information (e.g., an index represented by two bits) stored in the imaginary number portion.

[0033] The first weight table 222 outputs a basis vector b_w on a real axis corresponding to the index i_w, and the second weight table 224 outputs a basis vector b_p on an imaginary axis corresponding to the index i_p. In the first weight table 222, each element of a Walsh basis vector set is assigned an index (see FIG. 5A). In the second weight table 224, each element of a polar basis vector set is assigned an index (see FIG. 5B). The adder 230 sums the real basis vector b_w and the imaginary basis vector jb_p and outputs an antenna weight vector $[w_1, w_2, ..., w_L]$.

[0034] Briefly, the feedback information decoder 140 according to an embodiment of the present invention alternately stores feedback information in the real number portion and the imaginary number portion at feedback signaling intervals, sums the feedback information by way of sliding window, and obtains the antenna weight vector $[w_1, w_2, ..., w_L]$ based on the summed feedback information.

[0035] FIG. 3 is a block diagram of a receiving apparatus for transmission antenna diversity in a wireless communication system. Particularly, FIG. 3 shows an antenna weight measuring apparatus for measuring an antenna weight in the receiving apparatus.

[0036] Referring to FIG. 3, the receiving apparatus includes an antenna 300, a transmission antenna diversity channel information measuring unit 310, a basis vector transformer 320, an optimum weight detector 330, a feedback information uplink signal processor 340, and a data receiving processor 350. The data receiving processor 350 usually decodes a signal received through the antenna 300 and restores transmitting data.

[0037] The transmission antenna diversity channel information measuring unit 310 measures channel information from a signal received through the antenna 300 and outputs the result of measurement in the form of a matrix. The output channel information matrix is composed of L*M elements. "L" denotes the number of antennas, and "M" denotes the number of multi-path channels for each antenna. According to the description of FIG. 1, a transmitting apparatus transmits different pilot signals for discriminating antennas from one another to a receiving apparatus. The receiving apparatus measures each channel signal using a unique pilot signal corresponding to each of multiple antennas.

[0038] The basis vector transformer 320 transforms the channel information matrix output from the transmission antenna diversity channel information measuring unit 310 using a transform matrix composed of a complex basis vector set. The optimum weight detector 330 detects an element (i.e., a weight for maximizing a receiving SINR) of the complex basis vector set at which reception power with respect to the multiple antennas is maximum, using the transformed channel information matrix.

[0039] The feedback information uplink signal processor 340 transmits the result of detection as feedback information for controlling transmission antenna diversity to the transmitting apparatus through the antenna 300. Here, the feedback information uplink signal processor 340 makes the feedback information into a symbol according to a protocol suitable for feedback before transmitting the feedback information.

[0040] In a transmission antenna diversity method employing an SC method, it is essential to overcoming conventional problems to select an antenna having an optimum antenna weight among a plurality of transmitting antennas and to determine a form in which the optimum antenna weight is transmitted to the base station. Bearing this in mind, the operations of a mobile station will be described in detail. A mobile station (referred to as User Equipment (UE)) measures an optimum antenna weight from received channel information and feeds the result of measurement as feedback information back to a base station (referred to as UTRAN). Several embodiments will be described below.

[0041] In a first embodiment, a basis vector with respect to which maximum power is received is obtained using an orthonormal basis vector set and an index corresponding to the basis vector is fed back.

1) A channel information matrix H_{BW} obtained by transforming a receiving channel information matrix H using a transform matrix B_w composed of an or-

thonormal basis vector set is calculated as follows.

$$H_{BW} = HB_w$$

Here, $H = [h_1\ h_2\ h_3\ h_4]$, $B_w = [b_w(0)\ b_w(1)\ b_w(2)\ b_w(3)]$, h_1 is a column vector composed of multi-path channels transmitted from a first antenna, and $b_w(i)$ is a basis vector corresponding to an i-th index in the basis vector set. In the case of a binary system, the amount of calculation can be reduced by using a Hadamard matrix transform. In the other cases, the performance can be increased by using a high-speed algorithm suitable for the characteristics of each transform matrix.

2) A norm of each column vector constituting the channel information matrix H_{BW} is obtained. The values of norms are the value of receiving power measured with respect to the receiving channel information matrix H. The index of a basis vector corresponding to the maximum value among these values is the index of an orthonormal basis vector constituting an optimum weight.

3) The obtained index information is fed back to a base station (UTRAN). The above steps are repeated at each slot.

[0042] For example, when four transmitting antennas are used, 16 combinations of a complex basis vector are obtained from two basis vector sets. The above steps are performed on each complex basis vector, a complex basis vector showing maximum power is obtained, and the index of the complex basis vector is transmitted to the base station.

[0043] In a second embodiment, S vectors among M orthonormal basis vectors are used.

1) For example, 4 orthonormal basis vectors exist with respect to 4 antennas. Accordingly, S is one of 1 to 4. The values of M and S are stored.

2) Antenna selection weights $w_{b,i}$ are prepared from the result of step 1). The antenna selection weights $w_{b,i}$ are obtained by transforming antenna weights w_i used in a base station using a transform matrix B_w according to $w_{b,i} = B_w w_i$. The antenna selection weights $w_{b,i}$ for antennas other than a selected antenna are 0.

3) A measured reception power $P_i = w_{b,i}^H H_{BW}^H H_{BW} w_{b,i}$ is calculated using the antenna selection weight $w_{b,i}$ and a channel information matrix H_{BW}. Here, the channel information matrix H_{BW} is obtained by transforming a receiving channel information matrix H using a transform matrix B_w composed of an orthonormal basis vector set.

4) Steps 2) and 3) are repeated for ${}_MC_S$ cases, which is the number of cases in which S vectors can be selected from M basis vectors, ${}_MC_S = M!/\{(M-S)!\ S!\}$.

5) An antenna selection weight $w_{b,i}$ maximizing P_i in step 3) is selected.

6) The antenna selection weight $W_{b,i}$ obtained in step 5) is used as approximated feedback information.

[0044] For example, when four transmitting antenna are used, the number of combinations of a complex basis vector obtained from two basis vector sets is 16. When two antennas are selected according to an SC method, weights and power are obtained with respect to ${}_{16}C_2$ (=120) vector combinations, and then a combination of a basis vector maximizing the power is obtained. Antenna selection information and phase information as to the relative phase difference between the antennas may be included in feedback information.

[0045] In a third embodiment, a complex basis vector set is used, and the amount of feedback information is minimized. Here, a case of using four transmitting antennas will be explained as an example.

1) A channel information matrix H_{BW} obtained by transforming a receiving channel information matrix H using a transform matrix B_w composed of a Walsh basis vector set is calculated according to $H_{BW} = HB_w$. A channel information matrix H_{BP} obtained by transforming the receiving channel information matrix H using a transform matrix B_p composed of a polar basis vector set is calculated according to $H_{BP} = HB_P$. Here, $H = [h_1\ h_2\ h_3\ h_4]$, $B_w = [b_w(0)\ b_w(1)\ b_w(2)\ b_w(3)]$, $B_p = [b_p(0)\ b_p(1)\ b_p(2)\ b_p(3)]$, h_1 is a column vector composed of multi-path channels transmitted from a first antenna, $b_w(i)$ is a basis vector corresponding to an i-th index in the Walsh basis vector set, and $b_p(i)$ is a basis vector corresponding to an i-th index in the polar basis vector set.

2) It is assumed that $H_{BW}(i)$ indicates an i-th column vector in the matrix H_{BW}, and $H_{BP}(j)$ indicates a j-th column vector in the matrix H_{BP}. From this, a measured reception power $P_k(i,j) = \|H_{BW}(i)+jH_{BW}(j)\|^2$ is obtained when k = 0, 1, 2, ..., 15.
Here, I = 0, 1, 2, 3 and j = 0, 1, 2, 3.

3) Feedback information is generated based on k, i, and j which maximize the measured reception power. The above steps are repeated at each slot.

[0046] FIG. 4 is a block diagram of an embodiment of a receiving apparatus for transmission antenna diversity. The above third embodiment will be described with reference to FIG. 4. A receiving apparatus includes an antenna 400, a transmission antenna diversity channel information measuring unit 410, a basis vector transformer 420 having a Walsh basis vector transformer 422 and a polar basis vector transformer 424, an optimum weight detector 430 having first and second column adders 432 and 434, a combiner 436, a power calculator

438 and a maximum value detector 440, a feedback information uplink signal processor 450, and a data receiving processor 460. The data receiving processor 460 decodes a signal received through the antenna 400 and restores transmitting data.

[0047] The transmission antenna diversity channel information measuring unit 410 measures channel information from a signal received through the antenna 400 and outputs the result of measurement in the form of a matrix. The output channel information matrix H is composed of L*M elements. "L" denotes the number of antennas, and "M" denotes the number of multi-path channels for each antenna.

The Walsh basis vector transformer 422 transforms the channel information matrix H using a transform matrix composed of a Walsh complex basis vector set. The polar basis vector transformer 424 transforms the channel information matrix H using a transform matrix composed of a polar complex basis vector set.

[0048] The first column adder 432 sums the elements in all columns in each row in a matrix H_{BW} output from the Walsh basis vector transformer 422 and outputs a row vector $h_{BW}(i)$ expressed by

$$h_{BW}(i) = H_{BW}(i) \cdot 1_M.$$

Here, 1_M denotes a column vector whose length is M and whose elements are all 1.

[0049] The second column adder 434 sums the elements in all columns in each row in a matrix H_{BP} output from the polar basis vector transformer 424 and outputs a row vector $h_{BP}(j)$ expressed by

$$h_{BP}(j) = H_{BP}(j) \cdot 1_M.$$

[0050] The combiner 436 combines each of the row vectors h_{BW} with each of the row vectors h_{BP} and outputs a matrix H_B.

$$H_B(i, j) = h_{BW}(i) + jh_{BP}(j).$$

Here, i = 1, 2, ..., L and j = 1, 2, ..., L.

[0051] The power calculator 438 calculates the square of the modulus of each element of the combined matrix H_B and outputs a power matrix P_B.

$$P_B(i, j) = |H_B(i,j)|^2$$

Here, i = 1, 2, ..., L and j = 1 , 2, ..., L.

[0052] The maximum value detector 440 detects a maximum value from the power $P_B(i, j)$ with respect to each element and outputs an index (i_{max}, j_{max}) of an element corresponding to the maximum value.

$$(i_{max}, J_{max}) = \arg\max P_B(i,j)$$

[0053] The feedback information uplink signal processor 450 makes the index (i_{max}, j_{max}) to be transmitted to a transmitting apparatus into a symbol configured according to a protocol suitable for feedback and transmits the symbol through the antenna 400.

[0054] In a transmission antenna diversity method, as described above, a mobile station finds an optimum antenna weight through measurement of a channel. Here, a base station is required to send pilot signals discriminating antennas from one another for allowing the base station to measure a channel. To send different pilot signals for different antennas, a time division method, a frequency division method, or a code division method can be used. In the case of a Wideband Code Division Multiple Access (W-CDMA) standard, a method of using a multi-scrambling code, a multi-channelization code, or a multi-orthogonal pilot symbol pattern can be used to discriminate antennas from one another through pilot signals.

[0055] When two or more antennas are selected in a selection method, for efficient transmission, feedback information is transmitted in order of selection information and phase information from a mobile station to a base station. In other words, bit data corresponding to the selection information is sent first to allow relevant basis vectors to be selected, and then the phase information indicating the relation between basis vectors is sent. In the case of a protocol configured in frame unit, the selection information can be sent only in the first slot of a frame taking into account the property of a wireless fading environment in which the selection information scarcely changes but the phase information frequently changes. Here, when the selection information sent through the first slot of a frame is lost, this can influence the entire frame. Accordingly, it is preferable to perform error correction coding on selection information data before transmission. Besides such a case, the selection information and phase information can be selectively or both error correction coded before transmission.

[0056] When feedback information in transmitted in a plurality of slots for transmission from a mobile station (UE), the feedback information can be transmitted by way of progressively refining channel information every slot. In other words, only a portion of current data which is different from data previously fed back is transmitted. When the moving speed of the mobile station is high, the state of a channel may change while feedback information is fed back during an interval of a plurality of slots so that the state of a channel may not match the feedback information. In such conditions, the progressive refining method is effective. When simultaneously transmitting antenna selection information and phase information, it is preferable not to use a progressively refining mode of partially changing an FSM in order to prevent errors in the antenna selection information. Al-

though the entire antenna selection information and phase information fed back is not progressively refined, only the phase information can be conditionally refined according to the antenna selection information to improve the performance.

[0057] In the case of selecting three basis vectors from among four basis vectors to perform coherent phase correction in a four-transmission antenna diversity system, offset between selected antennas makes the imbalance of power among the antennas worse when the phase correction values of the selected antennas are the same. To overcome this problem, a first basis vector is weighted by 1/2, a second basis vector is weighted by one of exp(j*π/2+π/4), exp(j*2*π/2+π/4), exp(j*3*π/2+π/4), and exp(j*4*π/2+π/4), and a last basis vector is weighted by one of exp(j*π/2+π/8), exp(j*2*π/2+π/8), exp(j*3*π/2+π/8), and exp(j*4*π/2+π/8). Here, the number of cases of possible phases of each antenna is 4. Even when the number of antennas increases, the imbalance of power among the antennas can be minimized by rotating the constellation of phase correction values, which are multiplied by respective basis vectors, by a predetermined degree.

[0058] A transmitting apparatus and a receiving apparatus according to the present invention in a transmission antenna diversity system was described with reference to the attached drawings. As described above, a complex basis vector set is used for transmitting an optimum weight in a preferred embodiment of the present invention. Hereinafter, a case using a complex basis vector set will be described in detail.

[0059] A mobile station (UE) calculates an antenna weight to be applied to an access point of a transmitting antenna of a base station (UTRAN) in order to maximize reception power. For example, common pilot channels (CPICH) transmitted from four transmitting antennas are used for the calculation (see FIG. 1). When four transmitting antennas are used, an antenna weight is one complex basis vector included in a set of 16 complex basis vectors and is determined according to diversity employing an SC method. A real axis and an imaginary axis in a complex basis vector are composed of different orthonormal basis vectors. FIGS. 5A through 5C show examples of a basis vector set on the real axis, a basis vector set on the imaginary axis, and a complex basis vector set obtained by combining the two basis vector sets.

[0060] For each slot, the mobile station (UE) calculates an index corresponding to an optimum weight, that is, an index I corresponding to one complex basis vector selected from among the 16 complex basis vectors. Since basis vectors are alternately transmitted such that a real basis vector is transmitted in one slot and an imaginary basis vector is transmitted in the next slot, the index I transmitted at a time is one of numerals 0, 1, 2, 3, and 4 and is represented by EN-bit data. When an index is represented by a binary value I_{bin}, the relation between the binary value and I is expressed by

$$I_{bin} = \begin{cases} 00_{(2)}, & \text{if} \quad I=0 \\ 01_{(2)}, & \text{if} \quad I=1 \\ 10_{(2)}, & \text{if} \quad I=2 \\ 11_{(2)}, & \text{if} \quad I=3 \end{cases}$$

[0061] Here, I is a value used as an index value in FIGS. 5A and 5B showing the lists of orthonormal basis vectors. Each binary value I_{bin} is sequentially transmitted to the base station (UTRAN) through an FSM field. When $I_{bin} = 00_{(2)}$, 0 is sent as a most significant bit (MSB) and as a least significant bit (LSB). When $I_{bin} = 01_{(2)}$, 0 is sent as an MSB and 1 is sent as an LSB. When $I_{bin} = 10_{(2)}$, 1 is sent as an MSB and 0 is sent as an LSB. When $I_{bin} = 11_{(2)}$, 1 is sent as an MSB and as an LSB. Two-bit data is transmitted for a single slot time.

[0062] The base station (UTRAN) analyzes received feedback information according to FIG. 6. FIG. 6 shows the mapping relation between basis vectors b_w and b_p and the received feedback information (FSM) at each slot number. In FIG. 6, $b_w(i)$ is a vector corresponding to an i-th index in FIG. 5A, and $b_p(i)$ is a vector corresponding to an i-th index in FIG. 5B.

[0063] An antenna weight (a vector $w = w_{re} + jw_{im}$) calculated by the feedback information decoder 140 of the base station shown in FIG. 1 is a sliding window average of basis vectors received for the interval of two consecutive slots. The vector "w" is expressed by the following equation according to an algorithm.

$$\underline{w}(n) = \underline{w}_{re}(n) + j\underline{w}_{im}(n)$$

Here, $\underline{w}_{re}(n) = \underline{b}_w(2\lfloor n/2 \rfloor)$, and $\underline{w}_{im}(n) = \underline{b}_P(2\lfloor (n-1)/2 \rfloor)$.

[0064] FIG. 7 shows parameters and their values used for transmission antenna diversity employing a selection method using complex basis vectors when four antennas are used. In FIG. 7, in the case of a wireless protocol configured in a frame-slot structure as in the UMTS W-CDMA standard, a duration time in a slot is the time length of a single slot. The number of basis sets for basis rotation is the number of basis sets which are used. One basis vector set is used for a real axis, and another basis vector set is used for an imaginary axis. A feedback command length in slots is the number of slots occupied by a single command (information) to be used for determination of a weight. The number of selection index bits per signaling word is the number of bits necessary for representing selection information and is two when the number of antennas is four. The number of feedback information bits per slot is the number of bits of feedback information in a single slot. A feedback command update rate is an interval at which

feedback information is updated in a register of the base station. A feedback bit rate is information as to how many bits are fed back per second.

[0065] As described above, the present invention allows power to be equally distributed to transmitting antennas and maintains excellent performance at a high speed of movement, thereby minimizing the cost of configuring an RF processor. Particularly, by using information received in two consecutive slots, the present invention can be more reliably adapted to a channel at a low speed of movement. In addition, as for an extended selective combining method of selecting a plurality of antennas and coherently combining them, the present invention provides methods for improving the performance, thereby optimizing the performance. Accordingly, according to the present invention, hardware can be configured at a low cost, performance is excellent at a high speed of movement, and reliable channel adaptation can be accomplished at a low speed of movement, thereby maximizing channel capacity and link performance in a wireless mobile communication environment. The present invention can be applied to mobile communication systems such as CDMA-2000 systems and UMTS systems using a CDMA mode.

Claims

1. A closed loop transmission antenna diversity method employing a selective combining method when a plurality of antennas are used in a base station of a mobile communication system, the closed loop transmission antenna diversity method comprising the steps of:

   (a) measuring channel information from signals received through the plurality of antennas used in the base station and outputting a channel information matrix;
   (b) transforming the channel information matrix according to a transform matrix composed of a complex basis vector set;
   (c) calculating reception power with respect to the plurality of antennas based on the transformed channel information matrix; and
   (d) transmitting antenna selection information obtained based on the calculated reception power to the base station as feedback information for controlling transmission antenna diversity.

2. The closed loop transmission antenna diversity method of claim 1, wherein the step (a) comprises measuring channel information using pilot signals set differently for the plurality of antennas.

3. The closed loop transmission antenna diversity method of claim 1 or 2, wherein the step (b) comprises the sub steps of:

   (b1) calculating a first transformed channel information matrix from the channel information matrix using a transform matrix composed of a first basis vector set; and
   (b2) calculating a second transformed channel information matrix from the channel information matrix using a transform matrix composed of a second basis vector set, and
   the step (c) comprises the sub steps of:
   (c1) calculating reception power based on the first and second transformed channel information matrices; and
   (c2) detecting an element maximizing the reception power in the complex basis vector set.

4. The closed loop transmission antenna diversity method of claim 3, wherein the first and second basis vector sets are a Walsh basis vector set and a polar basis vector set, respectively.

5. The closed loop transmission antenna diversity method of any preceding claim, wherein the step (d) comprises alternately transmitting two indexes corresponding to a real part and an imaginary part, respectively, of a complex basis vector at feedback signaling intervals when an index corresponding to a basis vector included in the complex basis vector set is transmitted as the feedback information.

6. The closed loop transmission antenna diversity method of any preceding claim, wherein in the step (d) the feedback information comprises antenna selection information and phase information indicating a phase difference between antennas.

7. A closed loop transmission antenna diversity method employing a selective combining method, comprising the steps of:

   (a) receiving selection information related to a complex basis vector from a mobile station as feedback information for controlling transmission antenna diversity;
   (b) determining a complex basis vector selected based on the selection information;
   (c) obtaining an antenna weight for each antenna using the determined complex basis vector; and
   (d) generating a signal based on the antenna weight and transmitting the signal to the mobile station through a corresponding antenna.

8. The closed loop transmission antenna diversity method of claim 7, wherein the step (b) comprises the sub steps of:

(b1) receiving an index corresponding to an element of a complex basis vector set as the feedback information; and

(b2) selecting a complex basis vector corresponding to the index received in step (b1) by referring to a weight table in which an index is assigned to each element of a complex basis vector set composed of all combinations of first and second basis vector sets.

9. The closed loop transmission antenna diversity method of claim 7 or 8, wherein the step (a) comprises separately receiving as the feedback information the real part and imaginary part of an index corresponding to an element of a complex basis vector set for two feedback signaling intervals, and combining the real part and the imaginary part by way of sliding window.

10. The closed loop transmission antenna diversity method of claim 8, wherein the first and second basis vector sets are a Walsh basis vector set and a polar basis vector set, respectively.

11. A base station apparatus for a closed loop transmission antenna diversity method employing a selective combining method when a plurality of antennas are used in a mobile communication system, the base station apparatus comprising:

a plurality of antennas for receiving selection information related to a complex basis vector from a mobile station as feedback information for controlling transmission antenna diversity;

a feedback information decoder for determining a complex basis vector selected based on the selection information and obtaining an antenna weight for each antenna using the determined complex basis vector; and

a data transmitting unit for generating a signal based on the antenna weight and transmitting the signal to the mobile station through a corresponding antenna.

12. A mobile station apparatus for a closed loop transmission antenna diversity method employing a selective combining method when a plurality of antennas are used in a base station of a mobile communication system, the mobile station apparatus comprising:

a channel information measuring unit for measuring channel information from signals received through the plurality of antennas used in the base station and outputting a channel information matrix;

a basis vector transformer for transforming the channel information matrix according to a transform matrix composed of a complex basis vector set;

an optimum weight detector for calculating reception power with respect to the plurality of antennas based on the transformed channel information matrix and generating feedback information for controlling transmission antenna diversity based on the calculated reception power; and

an uplink signal processor for transmitting the feedback information to the base station in the form of a symbol configured according to a protocol suitable for feedback.

13. The mobile station apparatus of claim 12, wherein the basis vector transformer comprises:

a Walsh basis vector transformer for transforming the channel information matrix using a transform matrix composed of a Walsh basis vector set; and

a polar basis vector transformer for transforming the channel information matrix using a transform matrix composed of a polar basis vector set.

14. The mobile station apparatus of claim 12 or 13, wherein the optimum weight detector comprises:

first and second column adders each for adding elements in all columns in each row in the transformed channel information matrix and outputting a row vector;

a combiner for combining the outputs of the first and second column adders in all possible cases and outputting a combination matrix;

a power calculator for calculating power with respect to each element of the combination matrix; and

a maximum value detector for detecting a maximum value of the power with respect to each element and outputting an index of an element corresponding to the maximum value.

15. The mobile station apparatus of any of claims 12 to 14, wherein the uplink signal processor transmits antenna selection information and phase information as the feedback information.

FIG. 1

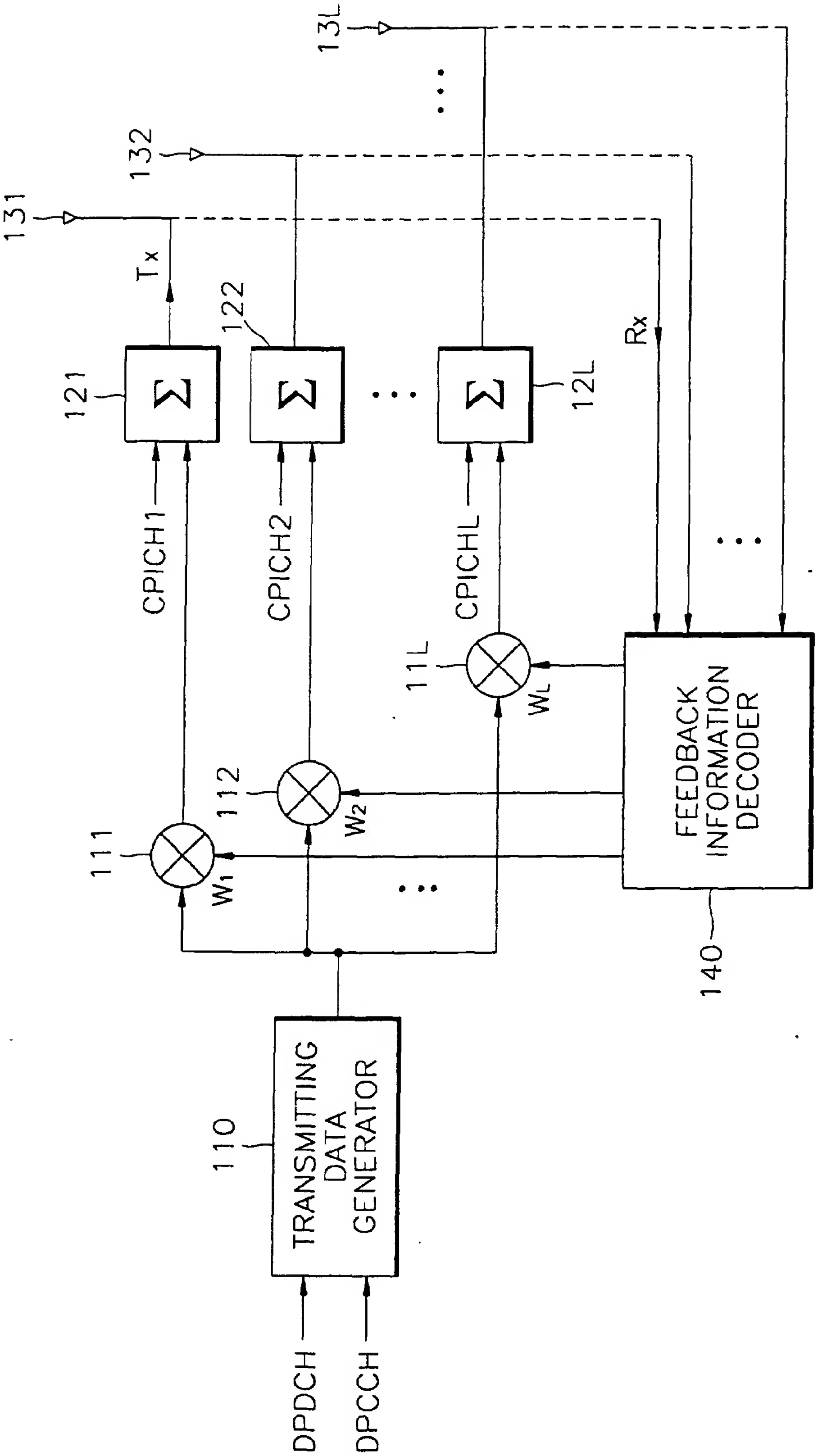
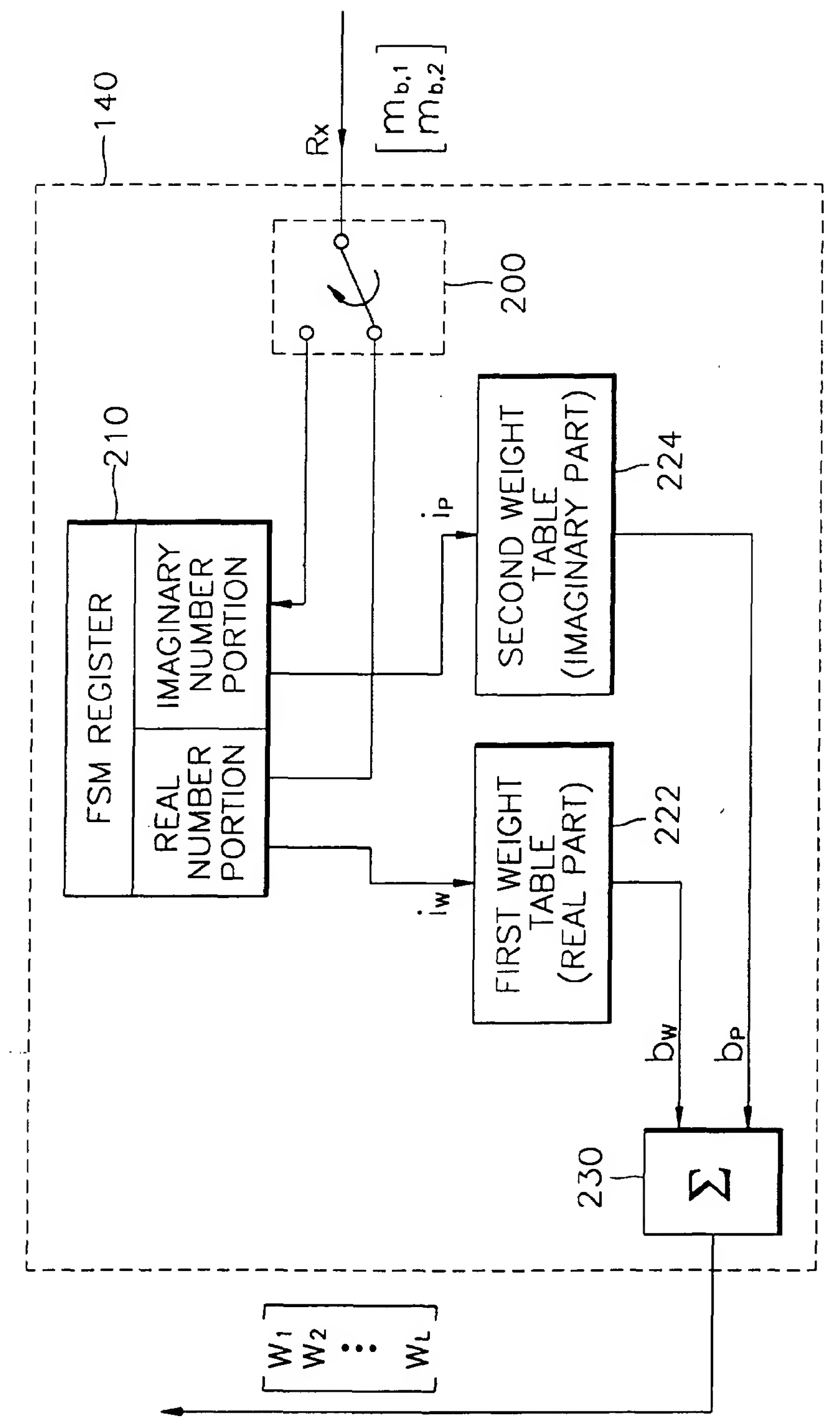

DPDCH
DPCCH
TRANSMITTING DATA GENERATOR
110
111
112
11L
W_1
W_2
W_L
CPICH1
CPICH2
CPICHL
121
122
12L
Tx
Rx
131
132
13L
FEEDBACK INFORMATION DECODER
140

FIG. 2



140
Rx
200
210
FSM REGISTER
REAL NUMBER PORTION
IMAGINARY NUMBER PORTION
FIRST WEIGHT TABLE (REAL PART)
SECOND WEIGHT TABLE (IMAGINARY PART)
222
224
230
Σ

FIG. 3

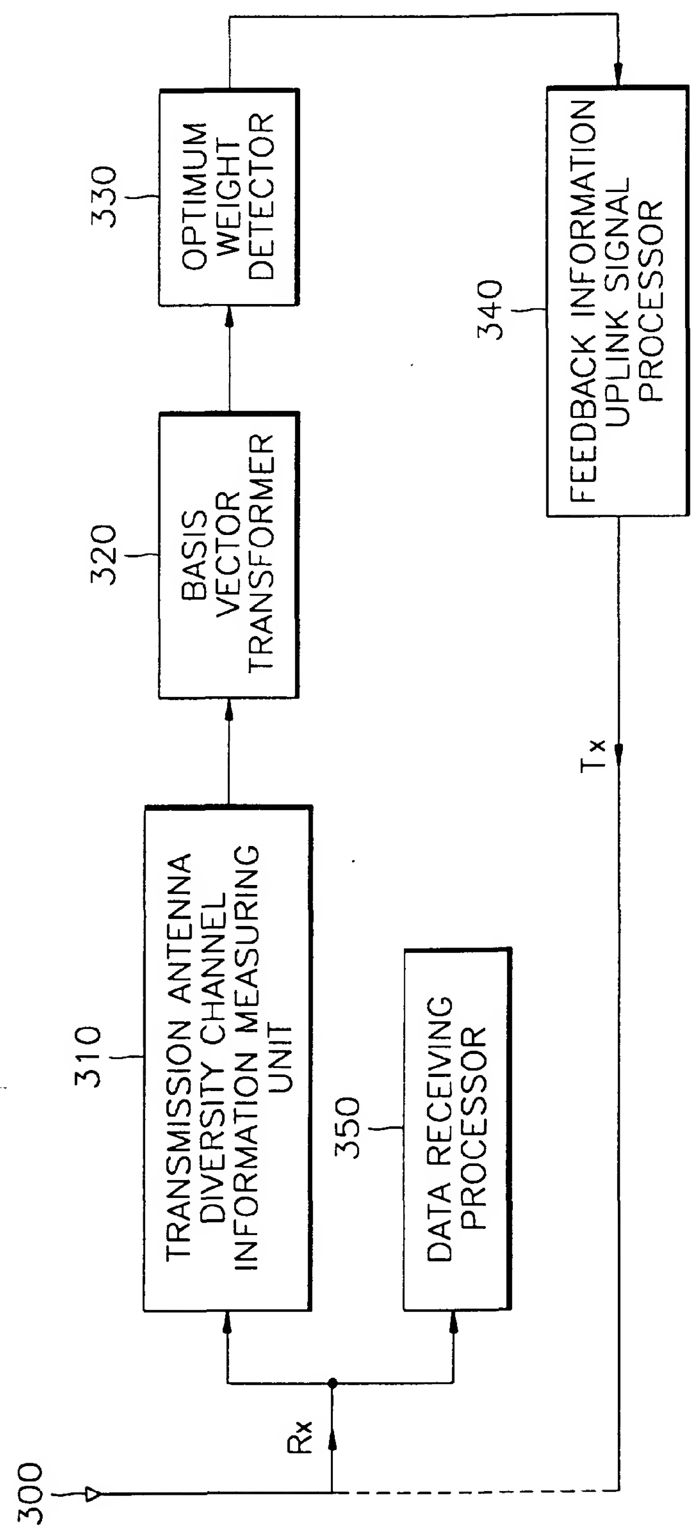

300
Rx
310
TRANSMISSION ANTENNA DIVERSITY CHANNEL INFORMATION MEASURING UNIT
320
BASIS VECTOR TRANSFORMER
330
OPTIMUM WEIGHT DETECTOR
340
FEEDBACK INFORMATION UPLINK SIGNAL PROCESSOR
Tx
350
DATA RECEIVING PROCESSOR

FIG. 4

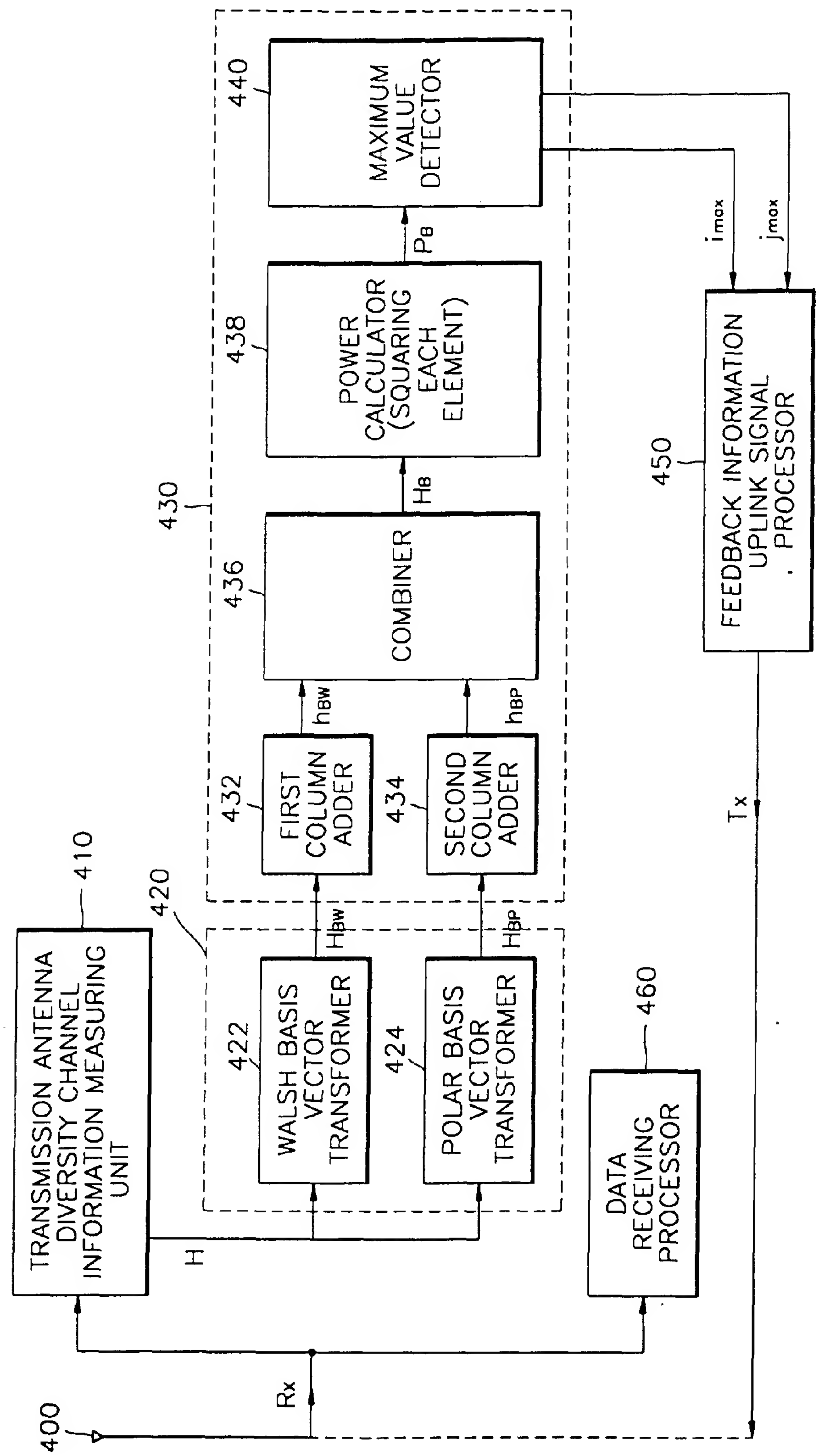

400
Rx
Tx
410
TRANSMISSION ANTENNA DIVERSITY CHANNEL INFORMATION MEASURING UNIT
H
420
422
WALSH BASIS VECTOR TRANSFORMER
424
POLAR BASIS VECTOR TRANSFORMER
HBW
HBP
430
432
FIRST COLUMN ADDER
434
SECOND COLUMN ADDER
hBW
hBP
436
COMBINER
HB
438
POWER CALCULATOR (SQUARING EACH ELEMENT)
PB
440
MAXIMUM VALUE DETECTOR
imax
jmax
450
FEEDBACK INFORMATION UPLINK SIGNAL PROCESSOR
460
DATA RECEIVING PROCESSOR

FIG. 5A

INDEX	VECTOR
0	[1 1 1 1]
1	[1 −1 1 −1]
2	[1 1 −1 −1]
3	[1 −1 −1 1]

FIG. 5B

INDEX	VECTOR
0	[−1 1 1 1]
1	[1 −1 1 1]
2	[1 1 −1 1]
3	[1 1 1 −1]

FIG. 5C

INDEX	VECTOR	INDEX	VECTOR
0	[1−j 1+j 1+j 1+j]	8	[1+j 1+j 1−j 1+j]
1	[1−j −1+j 1+j −1+j]	9	[1+j −1+j 1−j −1+j]
2	[1−j 1+j −1+j −1+j]	10	[1+j 1+j −1−j −1+j]
3	[1−j −1+j −1+j 1+j]	11	[1+j −1+j −1−j 1+j]
4	[1+j 1−j 1+j 1+j]	12	[1+j 1+j 1+j 1−j]
5	[1+j −1−j 1+j −1+j]	13	[1+j −1+j 1+j −1−j]
6	[1+j 1−j −1+j −1+j]	14	[1+j 1+j −1+j −1−j]
7	[1−j −1−j −1+j 1+j]	15	[1+j −1+j −1+j 1−j]

FIG. 6

SLOT NUMBER		0	1	2	3	• • •	14	15
FSM	00	$b_W(0)$	$b_P(0)$	$b_W(0)$	$b_P(0)$	• • •	$b_W(0)$	$b_P(0)$
	01	$b_W(1)$	$b_P(1)$	$b_W(1)$	$b_P(1)$	• • •	$b_W(1)$	$b_P(1)$
	10	$b_W(2)$	$b_P(2)$	$b_W(2)$	$b_P(2)$	• • •	$b_W(2)$	$b_P(2)$
	11	$b_W(3)$	$b_P(3)$	$b_W(3)$	$b_P(3)$	• • •	$b_W(3)$	$b_P(3)$

FIG. 7

PARAMETER	VALUE	TYPE
NUMBER OF ANTENNAS	$N_{ant} = 4$	CONSTANT
DURATION TIME IN A SLOT	$T_{slot} = 1/1500$ sec	
NUMBER OF BASIS SETS FOR BASIS ROTATION	$N_{set} = 2$	
FEEDBACK COMMAND LENGTH IN SLOTS	$N_W = 2$	
NUMBER OF SELECTION INDEX BITS PER SIGNALING WORD	$N_{set} = \log_2 N_{ant} = 2$	VARIABLE
NUMBER OF FEEDBACK INFORMATION BITS PER SLOT	$N_{FBD} = N_{set}/1 = 2$	
FEEDBACK COMMAND UPDATE RATE	$F_{up} = (N_{FBD}/N_W)T_{slot} = 1500Hz$	
FEEDBACK BIT RATE	$N_{FBD}/T_{slot} = 3000bps$	

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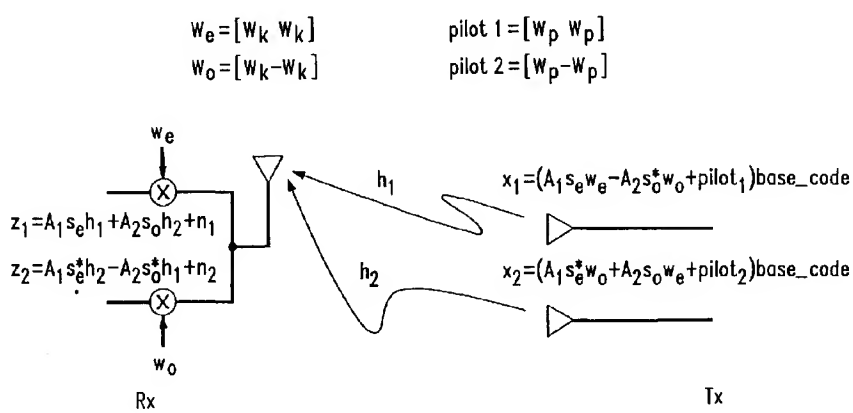
(54) Method for enhancing mobile cdma communications using space-time transmit diversity

(57) A smart antenna applies a space-time transmit diversity scheme to angular and temporary transmit diversity to combine with a polarization interleave switch beam to enhance the downlink wideband code division multiple access performance. The aperture gain is transferred to diversity gain by dividing a different angle of departure with different space-time spreading.

FIG. 12



EP 1 182 799 A2

Description

Field of the Invention

[0001] This invention relates to the field of telecommunications, and more particularly, this invention relates to the field of a Code Division Multiple Access telecommunications system using space-time transmit diversity.

Background of the Invention

[0002] The Universal Mobile Telecommunications Systems (UMTS), one of the third generation wireless proposals, is designed for use with various formats of voice/data services. The demand for data traffic implies-a higher throughput rate in the downlink traffic channels rather than in the uplink traffic channels. To improve the downlink capacity of a Wideband Code Division Multiple Access (WCDMA) system, studies and modifications of the physical layer structure of the UMTS in both phased array beamformers and the transmit diversity schemes have been accomplished.

[0003] A phased array beamforming approach forms a narrow transmitting beam directed to the desired mobile communication units. An aperture gain of the beamformer improves the downlink systems capacity. As the number of antenna elements is doubled, the coverage beamwidth in azimuth and the interference seen by mobiles is reduced to one-half. As a result, the downlink capacity is doubled. The downlink beamformer aims at the azimuth angle, which is estimated by the uplink processing of the desired mobile. The downlink beamformers can mis-aim in azimuth if downlink physical channels and uplink physical channels are different, e. g., in frequency division duplex (FDD) systems. additionally, downlink beamformers may perform worse than a single transmit antenna if the angle spread of the channels is larger than the 3 dB beamwidth of the beamformer.

[0004] To overcome some of these drawbacks, open-loop transmit diversity uses the space-time block code to transmit signals at two diversity antennas. Typically, the mobile can see at least two independent paths from the base station. One advantage of transmit diversity is that it creates diversity gain as compared to a single transmitting antenna. As a result, the received power can be above a certain level at the mobiles. In a slow and flat fading communications environment, the performance improvement of two units transmitting diversity over a single antenna can be as high as 7 dB. It has been found, however, that in a multiple time resolvable arrival environment, the performance improvement of two transmit diversity is diminished by the existing multipath diversity of a single antenna. Performance improvement can be as poor as 3 dB in slow fading channels and only 1 dB in fast fading channels. As is known, the full four-way transmit diversity does not exist. The throughput rate (bit rate) has to be rescued to three-fourths if four-way transmit diversity is achieved.

Summary of the Invention

[0005] The present invention is advantageous and provides a smart antenna, which applies a space-time transmit diversity (STTD) scheme to angular and temporary transmit diversity to combine with the polarization interleaved switch beam (SWB-ATTD) to enhance the downlink wideband Code Division Multiple Access (WCDMA) performance. The angular temporary transmit diversity and polarization interleaved switch beam is specially designed for multiple arrival and wide angle spread (urban) environments. In such an environment, the method of the invention transfers the aperture gain to diversity gain by dividing different angle of departure with different space-time spreading. The proposed switch beamformer preserves the aperture gain in time-resolvable arrival air interface, which is a typical WCDMA channel environment. It is possible to obtain an average 2-3 dB better than the steering space-time diversity and 6 dB superior achievement to single omni-transmit antenna in wide angle spread and rich multipath (urban) environments. The smart antenna of the present invention does not require per user's pilot and transceiver calibration at the base station for implementing beamformers. The interleaved polarization switch beam structure increases receiving diversity in the uplink traffic.

Brief Description of the Drawings

[0006] Other objects, features and advantages of the present invention will become apparent from the detailed description of the invention which follows, when considered in light of the accompanying drawings in which:

FIG. 1 is a chart showing the slot and frame structure for a downlink communication channel and the wideband CDMA system of UMTS.

FIG. 2 is a schematic block diagram showing the mixing of a channelization code and scrambling code for the UMTS and CDMA system of FIG. 1.

FIG. 3 shows a QPSK constellation for two antennas in a simple transmission scheme.

FIG. 4 is a block diagram showing the weights chosen to optimize the signal-to-noise ratio at the receiver of the system of FIG. 1.

FIG. 5 is a table showing the amplitude and phase of the weights of FIG. 4.

FIG. 6 is a table showing a transmission table for a single user using multi-carrier without an additional Walsh code.

FIG. 7 illustrates a transmission matrix with an extended Walsh code.

FIG. 8 is a transmission matrix of a 4x4 matrix with

three complex variables such as for multiple antennas.

FIG. 9 is a similar transmission matrix which guarantees orthogonality.

FIG. 10 is a transmission matrix having four-fold diversity prior to decoding.

FIG. 11 is a table showing the complexity difference between space-time diversity and OTD 1X receivers and operations per frame.

FIG. 12 is an overview of a space-time transmit diversity scheme.

FIG. 13 are equations showing diversity receiving at mobiles.

FIG. 14 is a block diagram showing a polarization interleave switch beam structure of the present invention.

FIG. 15 is a graph showing the combined space-time block code to antenna switch beams.

FIGS. 16 and 17 are tables showing performance analysis.

Detailed Description of the Preferred Embodiments

[0007] The present invention will now be described more fully hereinafter with reference to the accompanying drawings, in which preferred embodiments of the invention are shown. This invention may, however, be embodied in many different forms and should not be construed as limited to the embodiments set forth herein. Rather, these embodiments are provided so that this disclosure will be thorough and complete, and will fully convey the scope of the invention to those skilled in the art. Like numbers refer to like elements throughout.

[0008] The present invention is advantageous and provides a smart antenna, which applies a space-time transmit diversity (STTD) scheme to angular and temporary transmit diversity to combine with the polarization interleaved switch beam (SWB-ATTD) to enhance the downlink Wideband Code Division Multiple Access (WCDMA) performance. The angular temporary transmit diversity and polarization interleaved switch beam is specially designed for multiple arrival and wide angle spread (urban) environments. In such an environment, the method of the invention transfers the aperture gain to diversity gain by dividing different angle of departure with different space-time spreading. The proposed switch beamformer preserves the aperture gain in time-resolvable arrival air interface, which is a typical WCDMA channel environment. It is possible to obtain an average 2-3 dB better than the steering space-time diversity and 6 dB superior achievement to single omni-transmit antenna in wide angle spread and rich multipath (urban) environments. The smart antenna of the present invention does not require per user's pilot and transceiver calibration at the base station for implementing beamformers. The interleaved polarization switch beam structure increases receiving diversity in the uplink traffic.

[0009] For purposes of understanding, a brief description of transmit diversity applications in wideband code division multiple access systems and closed/open loop transmit diversity with space-time spreading Walsh code application and antenna options are first described.

[0010] The third generation cellular system, known as the Universal Mobile Telecommunications System (UMTS), is currently becoming standardized throughout Europe. It is designed to offer flexibility and advantages in wideband services over the present cellular systems, with various data rates as high as 2 Mb/s. UMTS is based on a Wideband Code Division Multiple access (WCDMA) physical layer structure. To improve capacity, antenna diversity has been addressed by may engineers. It has been found that for the purpose of low cost/size of a mobile terminal, receiver antenna arrays are not desirable in the downlink. It is possible to obtain the same performance gain by using multiple transmit antennas, and thus, transmit diversity (TD) has been used.

[0011] In UMTS, control channels are primarily encoded, while data channels are protected with a channel code. As known to those skilled in the art, it is possible to derive an expression for the bit-error rate (BER) in an encoded case.

[0012] In its typical application, UMTS is a wideband CDMA system with a five MHz bandwidth and 4.096 Mchip/s. It has variable spreading factors and can use multicode (several spreading codes are assigned to a user) to support variable bit-rates. The physical channels in the downlink are divided into one control channel **20,** and one data channel **22,** which are time-multiplexed. The slot and frame structure **24,26** for the downlink is shown in FIG. 1. The spreading is performed using a channelization code **28** and a scrambling code **30** (FIG. 2). The former is an Orthogonal Variable Spreading Factor (OVSF), which are Walsh-Hadamard codes of different lengths, and the latter is a 40960 chip (10 ms) segment of a $2^{18} - 1$ Gold code. The pulse shaping not shown in FIG. 2, is square-root raised cosine and the modulation is quadrature phase shift keying (QPSK).

[0013] There are currently two types of antenna transmit diversity (TD) proposals for UMTS: (1) open loop and (2) closed loop.

[0014] In the open loop scheme, the transmitter has no knowledge of the channels. To maximize the diversity, a space-time block code is used. This simple transmission scheme is shown in the table of FIG. 3, where $\square_1$, $\square_2$ belongs to a QPSK constellation. In a multipath channel with L resolvable paths, the received signal for the two symbols are:

r sub 1(t) = sum from {k=1} to L h sub {1,k} x sub 1 s(t-□sub k) + h sub {2,k} x sub 2 s (t - □sub k) + n sub 1 (t)

r sub 2(t) = sum from {k=1} to L - h sub {1,k} x * sub 2 s(t-□sub k) + h sub {2,k} x * sub 1 s (t - D sub k)

+ n sub 2 (t)

where s(t) is the spreading sequence. The outputs of rake finger k are:

$R_{1,k}$ = □r_1(t)s*(t-□$_k$)dt = $h_{1,k}$ x_1 + $h_{2,k}$ x_2 + $n_{1,k}$
$R_{2,k}$ = □r_2(t)s*(t-□$_k$)dt = - $h_{1,k}$ x^*_2 + $h_{2,k}$ x^*_1 + $n_{2,k}$

and maximum likelihood estimation becomes:

x_1 =
sum from {k=1} to L$h^*_{1,k}$ $r_{1,k}$ + $h_{2,k}$$r^*_{2,k}$ =
sum from {k=1} to L(□$h_{1,k}$□^{2} + □$h_{2,k}$□^{2})x_1 + $\tilde{n}_1$
x_2 =
sum from {k=1} to L$h^*_{2,k}$ $r_{1,k}$ - $h_{1,k}$$r^*_{2,k}$ =
sum from {k=1} to L(□$h_{1,k}$□^{2} + □$h_{2,k}$□^{2})x_2 + $\tilde{n}_2$

Hence, there will be 2L branch diversity. One advantage of this scheme is that the power between the two antennas is balanced, i.e. they are always the same.

[0015] In the closed loop transmit diversity scheme, the communications system transmits on both antennas **32,34** simultaneously, but with weights chosen to optimize the signal-to-noise ratio at the receiver (FIG. 4). Data **36** is forwarded with appropriate mixing with channelization code **28** and scrambling code **30.** The weights w_1 an w_2 are determined by the mobile and transmitted back to the base station. Three modes of feedback have been proposed with one, two and four bits respectively. These bits determine the amplitude and phase of the weights seen in the mode table of FIG. 5.

[0016] Mode 1 is selection diversity. The transmitter chooses the best antenna and transmits only on that one. Modes 2 and 3 are quantized version of the optimal weights. A special case of unquantized weights is feasible and assuming that the weights can be chosen with infinite precision, it is possible to calculate an analytical expression of the bit error rate (BER).

[0017] In a received signal with weighted transmission, the rake finger outputs and the maximum ratio combining (MRC) output can also be calculated.

r(t)=
sum from {k=1} to L$w_1$$h_{1,k}$ xs(t-□$_k$)+
sum from {k=1} to L$w_2$$h_{2,k}$ xs(t-□$_k$)+ n(t)
r_k = □r(t)s*(t-□$_k$) dt = $w_1$$h_{1,k}$x + $w_2$$h_{2,k}$x + n_k
x=
sum from {k=1} to L($w_1$$h_{1,k}$+$w_2$$h_{2,k}$)*$r_k$=
sum from {k=1} to L□$w_1$$h_{1,k}$+$w_2$$h_{2,k}$□^{2} x+ ($w_1$$h_{1,k}$+$w_2$$h_{2,k}$)$n_k$

[0018] The equivalent channel is h_k=$w_1$$h_{1,k}$+$w_2$$h_{2,k}$ and the objective is to choose the weights such that this channel has maximum power.

[0019] In order to estimate the channels, orthogonal pilot sequences are transmitted in every slot (even in mode 1). When there is a feedback bit rate of about 1600 bit/s, the weight update rate is 1600, 800 and 400 Hz, respectively, for modes 1, 2 and 3.

[0020] To analyze the performance, the following channel model is possible.

h(□;t)=
sum from {k=0} to {L-1} a_k(t)□(□-□$_k$)
where the path amplitudes ak(t) are independently fading. Furthermore, the channels for the two antennas are independent with the same average path powers,
E□□a_k□^{2}□, and delays, □$_k$. In the rake receiver, it can be assumed there is perfect dispreading, and any self interference can be ignored.

[0021] There will be a 2L path diversity for space-time transmit/diversty. The average signal to noise ratio on these paths are □$_1$, □$_1$, □$_2$, □$_2$, ..., □$_L$, □$_L$, assuming the two channels are identical on average. Furthermore, there is a 3 dB penalty because the transmit power is shared between the two antennas.

[0022] Orthogonal Transmit Diversity (OTD) is one method of obtaining downlink diversity without the drawbacks of mobile handset antenna diversity or multiple carrier or delay diversity. It takes an advantage of the decoding process when error correction codes are used. Orthogonal transmit diversity (OTD) is part of the current IS-2000 proposal known to those skilled in the art, and achieves diversity in the Viterbi decoder path metrics. OTD transmits alternate bits on different antennas. Even bits are transmitted on antenna 0 using one Walsh code, and odd bits are transmitted on antenna 1 using another Walsh code. In the IS-2000 standard, these codes are closely related. If, for example, a user *i* is assigned a Walsh code, $w_i^N(t)$, of length *N* in non-diversity mode, then the user *i* would be assigned two codes which are formed from $w_i^N(t)$ in the optional OTD mode. These two codes are formed as follows:

$w_i^{2N}(t) = [w_i^N(t)\ w_i^N(t)]$
$w_{i+N}^{2N}(t) = [w_i^N(t)\ -w_i^N(t)]$

where the code length has increased to 2N (reflected in the superscript) and there are now 2N possible codes (reflected in the subscript). Further, the second code $w_{i+N}^{2N}(t)$ is often referred to as the complementary code of $w_i^{2N}(t)$. Although the Walsh codes are extended, the overall data rate remains unchanged from the non-diversity mode. Each code or antenna carries one-half the original data.

[0023] By transmitting the even and odd data on different antennas, a form of diversity gain is obtained in Rayleigh fading conditions because the Viterbi decoder creates path metrics which are based on several consecutive bits after de-interleaving. Because alternate bits are transmitted from one of two antennas, the path metrics will inherently contain diversity. The diversity gain is a function of the strength of the code. The more powerful the code, the closer the performance will be to

a diversity scheme which obtains diversity on each symbol. At low Dopplers and for powerful codes (R=1/4 convolutional), the gains over no diversity can be dramatic. This scheme, however, is code dependent and low rate codes (R=1/2 convolutional) do not benefit as greatly from this scheme.

[0024] Space-time coding, on the other hand, can add diversity on the downlink without requiring an additional receive antenna, wasting bandwidth, or causing self-interference, and it is not dependent on the error correction code that is used. By coding over antennas and time in a particular way, diversity performance can be achieved without self-interference or extra bandwidth. This concept can be extended to the idea of Walsh coding, i.e., referred to as space-Walsh diversity or space-time spreading as used hereafter.

[0025] Space-time spreading may be readily implemented within the broad framework of the known IS-2000 system. Currently, the IS-2000 proposal supports two chip rates for signals, i.e., 1.2288 MHz (1X), and 3.6864 MHz (3X). The IX system is designed as a direct replacement for IS-95B systems, and has overlay capability. The 3X system supports higher data rates, and for the forward link only, supports a multiple carrier format, where each carrier has 1.2288 MHz chip rate. It also has overlay capability.

[0026] Currently, OTD is supported as an option within IS-2000. Thus, to make the inclusion of space-time spreading in IS-2000 straightforward, it is formulated in a manner which is similar to OTD.

[0027] Using the OTD framework specified in the IS-2000 proposal, space-time spreading can be readily applied to the system. Space-time spreading would require the sharing of spreading codes (i.e., two users share the use of two Walsh codes). Implementing space-time spreading in this fashion, however, is undesirable. First, the sharing of codes could cause problems if the transmit powers to the two users sharing Walsh codes were radically different. In such a case, imperfect channel estimation could result in significant cross terms. Additionally, OTD uses extended Walsh codes, which eliminate the need for code sharing and maximize the commonality with OTD within the standard. This requires fewer changes to the standard.

[0028] One main difference between space-time spreading and OTD is in the mapping of data onto the transmit diversity antennas.

[0029] As noted before, space-time spreading is readily implemented with few changes to an OTD framework. Because each user is assigned two extended Walsh codes (formed from a single non-extended Walsh code), and the data is partitioned into two streams, the diversity scheme may be applied to the two streams as if they were two different users. On the first antenna, the system would transmit:

$$x_1(t) = \square P/2\ [s_e(t)w(t)-s_o(t)^*\square(t)]\ p(t)$$

where P represents the total transmit power, $s_e(t) = Y_{I1}+Y_{Q1}$ is the even symbol stream, and $s_o(t) = Y_{I2}+Y_{Q2}$ is odd symbol stream. The Walsh code w(t) and its complement are used to spread the signal and are extended Walsh codes.

[0030] More importantly, it is possible to provide four-fold diversity at the encoded symbol level by assigning an extra Walsh code to a user, which will not cause cross-interference with other users regardless of channel estimation accuracy. These gains could be significant for users which have low I_{OR}/I_{OC}[7] values. Ordinarily, these users would be in a soft handoff. If the resources do not exist in the adjacent cell to support this user, then the use of additional Walsh resources on a per user basis may be appropriate.

[0031] If sacrificing an additional Walsh code is not desirable, it is possible to obtain diversity prior to the decoder in a multi-carrier system. Each symbol is sent four times (thus four-fold diversity). Instead of using an extra code, a single Walsh code is extended and each symbol sent twice. Each data stream s_i would then be split into two data streams $s^e{}_i$ and $s^o{}_i$ (even and odd). Instead of transmitting each signal four times, it is possible to transmit each signal twice, thus obtaining two-fold diversity, as shown in the table of FIG. 6. Equivalently, it is possible to use the transmission matrix notation developed to yield the transmission matrix of FIG. 7. The transmission matrix is an established type of matrix. The rows can represent Walsh codes (orthogonal channels) (non-orthogonal channels).

[0032] There are many possible assignments of bits to corresponding Walsh codes, antennas and carriers. Each will probably not differ significantly in terms of performance.

[0033] This transmission scheme will lead to a combining method, which is similar to the four-fold diversity case with the exception that only two Walsh outputs are used per decision statistic, and it will only yield a two-fold diversity improvement.

[0034] Unfortunately, a full-rate orthogonal design for a four antenna/channel system with complex signaling (e.g., QPSK modulation 1X system) does not exist. It is possible, however, to to obtain diversity improvement using space-time spreading with four antennas.

[0035] A modified transmission matrix for three antennas is possible. To allow four transmit antennas, the Walsh code is extended for a particular user twice, to obtain four Walsh codes with four times the length.

[0036] With a four-fold matrix extension, it is possible to have no code sharing. To obtain a transmission matrix, an orthogonal matrix with four columns and thus at least four rows is required. While a 4x4 orthogonal matrix with four complex variables does not exist, a 4x4 matrix with three complex variables does exist. One such transmission matrix is shown in FIG. 8.

[0037] For the matrix in FIG. 8, $H(t)H = (\square h_1\square^2 + \square h_2\square^2 + \square h_3\square^2 + \square h_4\square^2\ I$. This achieves four-fold diversity. In order to achieve this, however, the data rate must

be reduced to three-fourths the original rate. This can be seen by noticing that while it is possible to use four codes (i.e., the rows of T), it is only possible to transmit three symbols.

[0038] A second option for using four transmit antennas without reducing the data rate is to use the transmission matrix as shown in FIG. 9.

[0039] This matrix guarantees orthogonality, but only achieves two-fold diversity before the decoder. If the interleaving is done correctly, however, the metrics are:

$$b_1(|h_1|^2 + |h_2|^2), b_2(|h^3|^2 + |h_4|^2), b_3(|h_1|^2 + |h_2|^2),$$
$$b_4(|h_3|^2 + |h_4|^2), \ldots$$

[0040] While two-fold diversity is achieved prior to decoding, the Viterbi decoder can see up to four-fold diversity due to the path metrics. The decoder achieves the diversity gain from two to four, without data rate loss. As in OTD, the gains are dependent on the strength of the error correction code.

[0041] The last option is similar to option 1 and uses an orthogonal design. The transmisson matrix is shown in FIG. 10. This option also achieves four-fold diversity prior to decoding, but also suffers from a 25% loss in data rate. This allows all four codes to be used on all four antennas.

[0042] A major difference between OTD and space-time spreading in the mobile receiver is in the baseband receiver functions. All other components remain the same. The baseband receiver performs complex code uncovering, Walsh despreading, channel estimation, channel compensation, multiplexing/de-interleaving, and Viterbi decoding.

[0043] This complex uncovering requires a multiplication for each sample of the I&Q streams for all fingers for each of the early, late and on-time samples. Assuming eight samples per chip and three fingers, this results in approximately $3.5 \cdot 10^6$ operations per frame. The Walsh uncovering then requires approximately $9 \cdot 10^5$ operations per frame. Neither of these operation counts depends on the diversity scheme used.

[0044] Modeling channel estimation as a simple PCG average for each of two Walsh codes (one per transmit antenna) results in approximately $9 \cdot 10^5$ operations per frame. Channel compensation will depend on the data rate and requires about $5.5 \cdot 10^4$ N operations per frame for OTD, where N is a scale factor that increases with data rate (N=1 for voice). Space-time spreading is twice as complex as OTD in this one area.

[0045] The Viterbi decoding complexity also depends on the data rate and code rate. For RC4 the Viterbi decoding requires roughly $1.1 \cdot 10^6$ N operations pre frame. In addition to the detection and decoding functions, the receiver must perform searcher functions to track multipath. The searcher requires $16 \cdot 10^6$ operations per frame. The total operation count for both OTD and space-time spreading at data rates of 9.6 kbps and 76.8 kbps are shown in the table of FIG. 11. As can be seen, the complexity increase (0.2% for voice, 1.4% for 76.8 kbps) is extremely minor.

[0046] Space-time spreading offers significant performance gains over OTD. Space-time spreading outperforms OTD for all cases, primarily because it provides diversity prior to the decoding process while OTD relies on the Viterbi decoder. For R=1/2 convolutional codes at low speed, the performance gains are large (5 dB to 8 dB) for fundamental channels. For R=1/2 convolutional codes for all other speeds, the performance gains are significant for fundamental, supplemental and common channels (1 dB to 3 dB). Further, the gains of space-time spreading over no diversity are dramatic while the gains of OTD over no diversity are moderate to small with rate 1/2 coding. The performance gains of space-time spreading over OTD are more modest when 1/4 rate coding is used with gains ranging from 0.3dB to 0.7dB. Both space-time spreading and OTD offer large gains over no transmit diversity.

[0047] Space-time spreading offers more flexibility in choice of radio configuration. STS makes the capacity of RC3 with OTD (R=1/4) and RC4 with STS (R=1/2) approximately equal. Additionally, space-time spreading does not introduce significant computational complexity and both space-time spreading and OTD can be supported in the mobile and base station with relatively minor adjustments.

[0048] The present invention uses the advantages of space-time spreading. In accordance with the present invention, a smart antenna is based on the switch beam and space-time spreading. Space-time spreading implements the space-time block code to different transmit antennas, which have potentially independent channels to the mobile. The signals transmitted at antenna x_1 and x_2 can be written as follows:

$$x_1=(A_1 s_e w_e - A_2 s^*_o w_o + pilot_1)base\_code$$
$$x_2=(A_1 s^*_e w_o + A_2 s_o w_e + pilot_2)base\_code$$

where A and pilot denote amplitude and common pilot channel, s_e and s_o denote even and odd bits in bit sequence, and w_e and w_o denotes the mutually orthogonal extended Walsh code for even and odd bits. That is:

$$w_e = [\, w_k \; w_k \,]$$
$$w_o = [\, w_k \; -w_k \,]$$

where w_k is the Walsh code for the *k*th user. The transmitted signals are summed up in the air. The received signals are separately multiplied by the extended Walsh code w_e and w_o to obtain z_1 and z_2 a shown in FIG. 12. The above two outputs can be combined by applying the channel coefficients, which are estimated from common pilots, as shown in FIG. 13.

[0049] FIG. 14 shows the proposed antenna structure for polarization interleaved switch beam with angular and temporary transmit diversity. A positive and nega-

tive **45□** oriented polarization phased array **50** is implemented at the tower top. There are a total of eight branches **52** of antennas: four branches for +45□ polarization, and four branches for -45□ polarization. Four cables are required by implementing polarization interleaved switch beams so there are only four I/O ports **54** at the base station for each sector. The Butler matrix **56** is divided into two stages and sandwiches the power amplifiers **58**. This structure not only reduces the insertion loss of the Butler matrix after power amplifiers, but evenly loads the power amplifiers for a specific switch beam.

[0050] FIG. 15 shows the arrangement of the space-time spreading and the common pilot tone of the smart antenna shown in FIG. 14. The overlapped region for any adjacent beams is 3 dB lower from the top of the beam. Each adjacent beam has different polarization so polarization diversity gain will be observed in the overlapped region. The interleaved polarization structure does not improve the downlink approach. The adjacent beams are orthogonal to each other by applying the space-time block code.

[0051] The chip rate of UMTS is very high (3.86 Mchip per second) so the relative arrival delay of each propagation ray can exceed one chip's duration. Such a multiple arrival in time domain creates diversity receiving at mobiles even though one transmit antenna is used. Therefore, adding another transmit antenna/diversity does not have significant performance gain as aforementioned (1-3 dB). On the other hand, beamforming gain is not affected by multipath delay. But beamforming gain can be reduced by the propagation angle spread observed at the base stations.

[0052] The steering space-time transmit diversity scheme of the present invention improves the downlink capacity when four branches of transmitting antennas are used. Two groups of two-element phased array (□/2 spacing) are placed 10 □ apart to have a 3 dB beamformer and two-way spatial diversity. The beamforming gain can be reduced in wide angle spread because of mismatch in the beamformer's aiming angle as well as distortion in the pilot tone's phase. The ST-STTD creates two-way spatial diversity at all times with an occasionally maximum 3 dB beamforming gain, which can be reduced by pilot and aiming mismatch. The diversity transmitting has higher priority than the beamformer in this scheme.

[0053] The polarization interleaved switch beam with angular and temporary transmit diversity method of the present invention combines two-way transmit diversity with different polarization interleaved switch beam. The beamforming gain is equal to 6 dB divided by the number of beam implemented to fulfill the angle spread. The diversity comes from the different angle of departure plus the antenna polarization. Polarization interleaved switch beam with angular and temporary transmit diversity preserves the aperture gain as best it can while the transmit diversity does not exist at all time. The beamformer gain is placed at higher priority than the transmit diversity gain in this method. Because the multipath arrival in WCDMA can reduce the transmit diversity gain over one antenna, this method has better performance than the steering space-time transmit diversity method. Performance analysis is shown in Tables 1 and 2 (FIGS. 15 and 16).

[0054] Many modifications and other embodiments of the invention will come to the mind of one skilled in the art having the benefit of the teachings presented in the foregoing descriptions and the associated drawings. Therefore, it is to be understood that the invention is not to be limited to the specific embodiments disclosed, and that the modifications and embodiments are intended to be included within the scope of the dependent claims.

Claims

1. A method for enhancing the communication signal within a wideband code division multiple access mobile communications system using space-time diversity comprising the step of:

   transferring the operative gain of an antenna structure to diversity gain by dividing a different angle of departure with different space-time spreading.

2. A method for enhancing the communication signal within a wideband code division multiple access communications system using space-time diversity comprising the step of:

   transferring the operative gain of a switch beamformer antenna structure to diversity gain by dividing a different angle of departure with different space-time spreading; and
   mismatching an aiming angle of the switch beamformer antenna structure and distorting the phase of a pilot tone for reducing beamformer gain within a side angle spread.

3. A method according to Claim 1 or 2, and comprising the step of imparting a spreading function to the communications signal using mutually orthogonal extended Walsh codes for even and odd bits of data.

4. A method according to Claim 3, wherein the extended Walsh codes are formed from a single non-extended Walsh code.

5. A method according to Claim 3, and comprising the step of receiving the communication signal within a mobile communications unit and separately multiplying the received communication signal by the extended Walsh codes.

6. A method according to Claim 5, and comprising the step of estimating channel coefficients from common pilot signals and combining the output of the multiplying of the received signals by applying the channel coefficients.

7. A method according to Claim 5, and comprising the step of estimating channel coefficients from common pilot signals and combining the output of the previous multiplication step of the received signals by applying the channel coefficients.

8. A method according to Claim 1, and comprising the step of preserving the aperture gain in a time-resolvable arrival air interface with a switch beamformer antenna structure.

9. A method according to Claim 2, and comprising the step of preserving the aperture gain in a time-resolvable arrival air interface with the switch beamformer antenna structure.

10. A method according to Claim 2 or 8, and comprising the step of using four branches of the switch beam former antenna structure with two groups of two-element phased array and two-way spatial diversity.

11. A method according to Claim 10, and comprising the step of mismatching a beamformer's aiming angle and distorting the phase of a pilot tone for reducing beamforming gain within a side angle spread.

12. A method according to Claim 2 or 10, and comprising the step of placing beam former gain at higher priority than transmit diversity gain.

FIG. 1

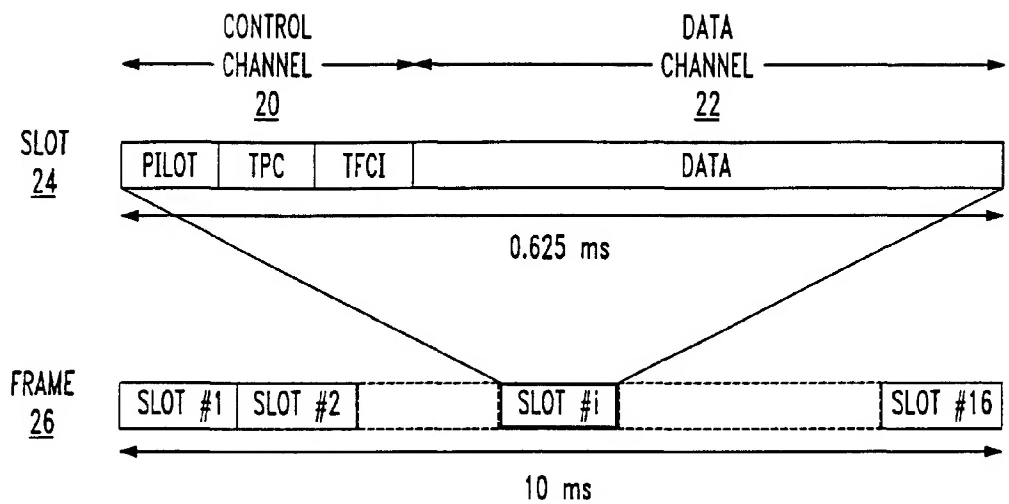

CONTROL
CHANNEL
20
DATA
CHANNEL
22
SLOT
24
PILOT
TPC
TFCI
DATA
0.625 ms
FRAME
26
SLOT #1
SLOT #2
SLOT #i
SLOT #16
10 ms

FIG. 2

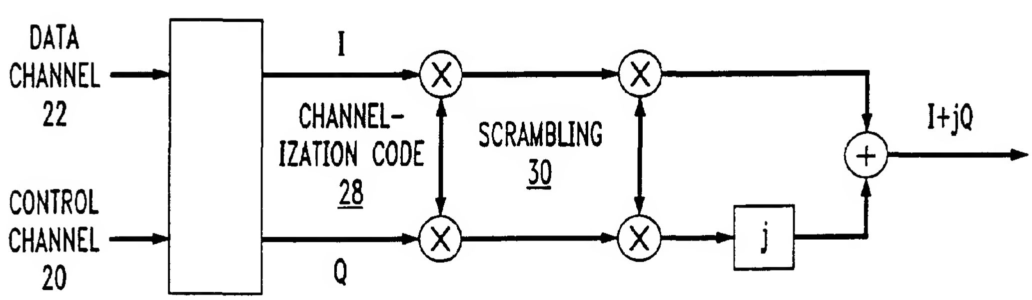

DATA
CHANNEL
22
CONTROL
CHANNEL
20
I
Q
CHANNEL-
IZATION CODE
28
SCRAMBLING
30
j
I+jQ

FIG. 3

	ANTENNA 1	ANTENNA 2
SYMBOL 1	x_1	x_2
SYMBOL 2	$-x_2^*$	x_1^*

SPACE–TIME TD

FIG. 4

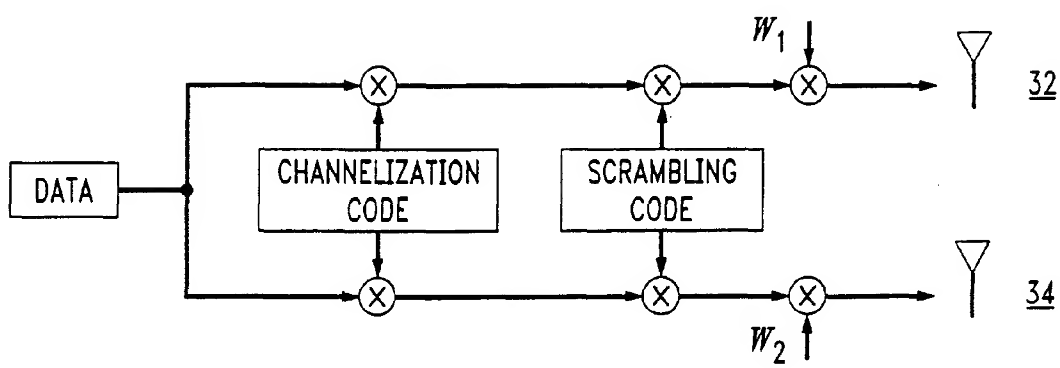


FIG. 5

MODE	POWER		PHASE	
	BITS	VALUES	BITS	VALUES
1	1	0, 1	0	–
2	0	0.5	2	$0, \frac{\pi}{2}, \pi, \frac{3\pi}{2}$
3	1	0.2, 0.8	3	$0, \frac{\pi}{4}, \frac{\pi}{2}, \frac{3\pi}{4}, \pi, \frac{5\pi}{4}, \frac{3\pi}{2}, \frac{7\pi}{4}$

FIG. 6

ANTENNA	FREQUENCY BAND	w'_1	$\overline{w'_1}$
1	$f1$	s_1^e	s_2^e
1	$f2$	s_2^o	s_3^e
1	$f3$	s_3^o	s_1^o
2	$f1$	$(-s_2^e)^*$	$(s_1^e)^*$
2	$f2$	$(-s_3^e)^*$	$(s_2^o)^*$
2	$f3$	$(-s_1^o)^*$	$(s_3^o)^*$

TRANSMISSION TABLE FOR A SINGLE USER USING MULTI-CARRIER WITHOUT ADDITIONAL WALSH CODE

FIG. 7

$$T = \begin{bmatrix} s_1^e & (-s_2^e)^* \\ s_2^o & (-s_3^e)^* \\ s_3^o & (-s_1^o)^* \\ s_2^e & (s_1^e)^* \\ s_3^e & (s_2^o)^* \\ s_1^o & (s_3^o)^* \end{bmatrix}$$

FIG. 8

$$T = \begin{bmatrix} b_1 & -b_2 & 0 & b_3^* \\ b_2^* & b_1^* & -b_3^* & 0 \\ 0 & b_3 & b_1 & b_2^* \\ b_3 & 0 & b_2 & -b_1^* \end{bmatrix}$$

FIG. 9

$$T = \begin{bmatrix} b_1 & -b_2 & 0 & 0 \\ 0 & 0 & b_3 & -b_4^* \\ b_2^* & b_1^* & 0 & 0 \\ 0 & 0 & b_4^* & b_3^* \end{bmatrix}$$

FIG. 10

$$T = \begin{bmatrix} b_1 & b_2 & \frac{b_3}{\sqrt{2}} & \frac{b_3}{\sqrt{2}} \\ -b_2^* & b_1^* & \frac{b_3}{\sqrt{2}} & -\frac{b_3}{\sqrt{2}} \\ \frac{b_3^*}{\sqrt{2}} & \frac{b_3^*}{\sqrt{2}} & \frac{-b_1-b_1^*+b_2-b_2^*}{2} & \frac{-b_2-b_2^*+b_1-b_1^*}{2} \\ \frac{b_3^*}{\sqrt{2}} & \frac{-b_3^*}{\sqrt{2}} & \frac{b_1-b_1^*+b_2-b_2^*}{2} & \frac{-b_2+b_2^*-b_1-b_1^*}{2} \end{bmatrix}$$

FIG. 11

Diversity Scheme	9.6kbps	76.8kbps
OTD	$22.46 \cdot 10^6$	$30.54 \cdot 10^6$
STS	$22.51 \cdot 10^6$	$30.98 \cdot 10^6$
% increase	0.2 %	1.4 %

FIG. 12

$W_e = [W_k \ W_k]$
$W_o = [W_k - W_k]$

pilot $1 = [W_p \ W_p]$
pilot $2 = [W_p - W_p]$

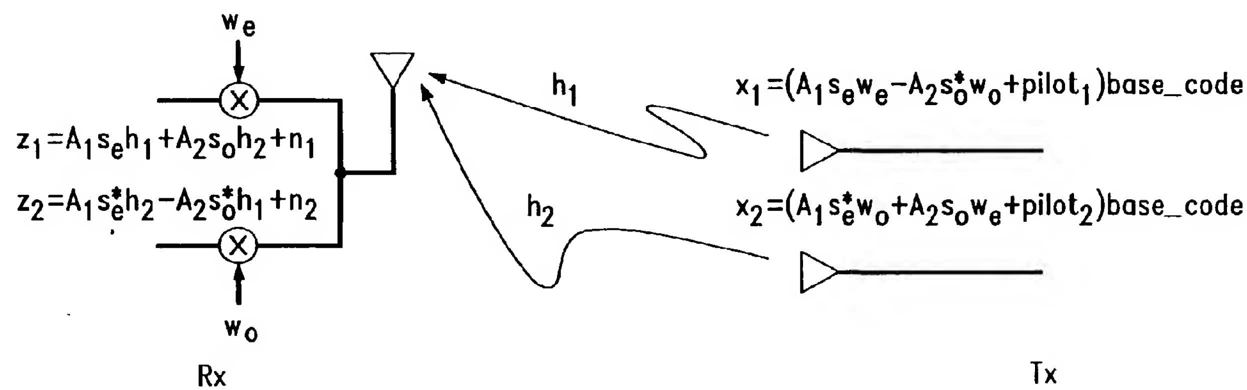


FIG. 13

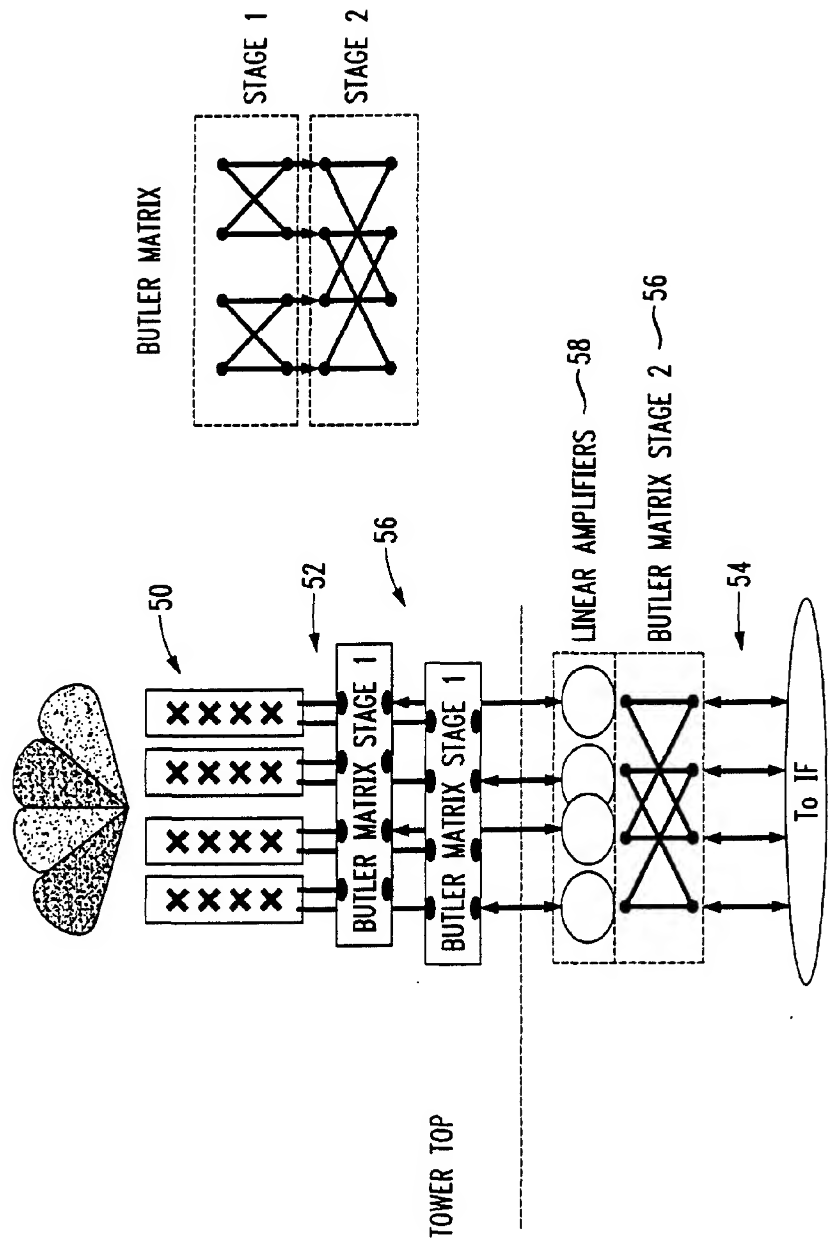
$$[\hat{s}_e \ \ \hat{s}_o] = \underbrace{\boxed{\left([A_1 s_e \ \ A_2 s_o] \cdot \begin{bmatrix} h_1 & h_2^* \\ h_2 & -h_1^* \end{bmatrix} + [n_1 \ \ n_2] \right)}}_{[z_1 \ \ z_2^*]} \cdot \begin{bmatrix} \hat{h}_1^* & \hat{h}_2^* \\ \hat{h}_2 & -\hat{h}_1 \end{bmatrix}$$

THE ABOVE EQUATION CAN BE REWRITTEN AS,

$$\begin{bmatrix} \hat{s}_e \\ \hat{s}_o \end{bmatrix} = \begin{bmatrix} A_1 s_e(|h_1|^2 + |h_2|^2) \\ A_2 s_o(|h_1|^2 + |h_2|^2) \end{bmatrix} + [\tilde{n}_1 \ \ \tilde{n}_2]$$

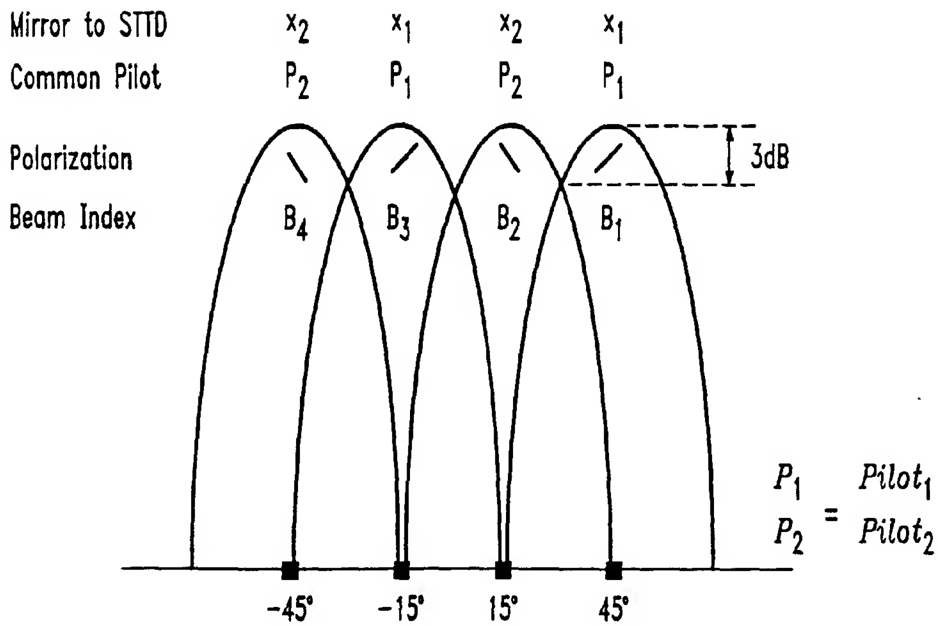
WHICH SHOWS THE DIVERSITY RECEIVING AT THE MOBILES.

FIG. 14



THE POLARIZATION INTERLEAVED SWITCH BEAM STRUCTURE

FIG. 15



Combine space-time block code to switch beams.

FIG. 16

Air Interface \ Angle Spread		LOS prop.	<15°	30°	45°	>60° (2 rays)
No time resolvable path (flat fading)	Slow fading	3-6 (6)	6-9 (10)	8-10 (7)	8-10 (7)	7-10 (7)
	Fast fading	3-6 (6)	4-6 (5)	5-6 (3)	5-6 (2)	2-5 (2)
2 equal power time resolvable paths	Slow fading		5-6 (6)	6 (4)	6 (3)	3-5 (3)
	Fast fading		3-6 (4)	4-6 (2)	4-5 (1)	1-4 (1)

Table 1. SWB_ATTD downlink performance gain over 1 Tx antenna, () denote the gain of ST-STTD

FIG. 17

Air Interface \ Angle Spread		LOS prop.	<15°	30°	45°	>60° (2 rays)
No time resolvable path (flat fading)	Slow fading	3 (3)	-4-2 (3)	2-3 (5)	6 (6)	6 (6)
	Fast fading	3 (3)	1-3 (3)	2-4 (4)	4 (4)	4 (4)
2 equal power time resolvable paths	Slow fading		0-3 (3)	2-4 (4)	5 (5)	5 (5)
	Fast fading		2-3 (3)	(3)	3 (3)	3 (3)

Table 2. Uplink performance gain over 2 Rx antenna, () denotes the gain of ST-STTD

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(54) Adaptive time deversity and spatial diversity for OFDM

(57) An adaptable orthogonal frequency-division multiplexing system (OFDM) that uses a multiple input multiple output (MIMO) to having OFDM signals transmitted either in accordance with time diversity to reducing signal fading or in accordance with spatial diversity to increase the data rate. Sub-carriers are classified for spatial diversity transmission or for time diversity transmission based on the result of a comparison between threshold values and at least one of three criteria. The criteria includes a calculation of a smallest eigen value of a frequency channel response matrix and a smallest element of a diagonal of the matrix and a ratio of the largest and smallest eigen values of the matrix.

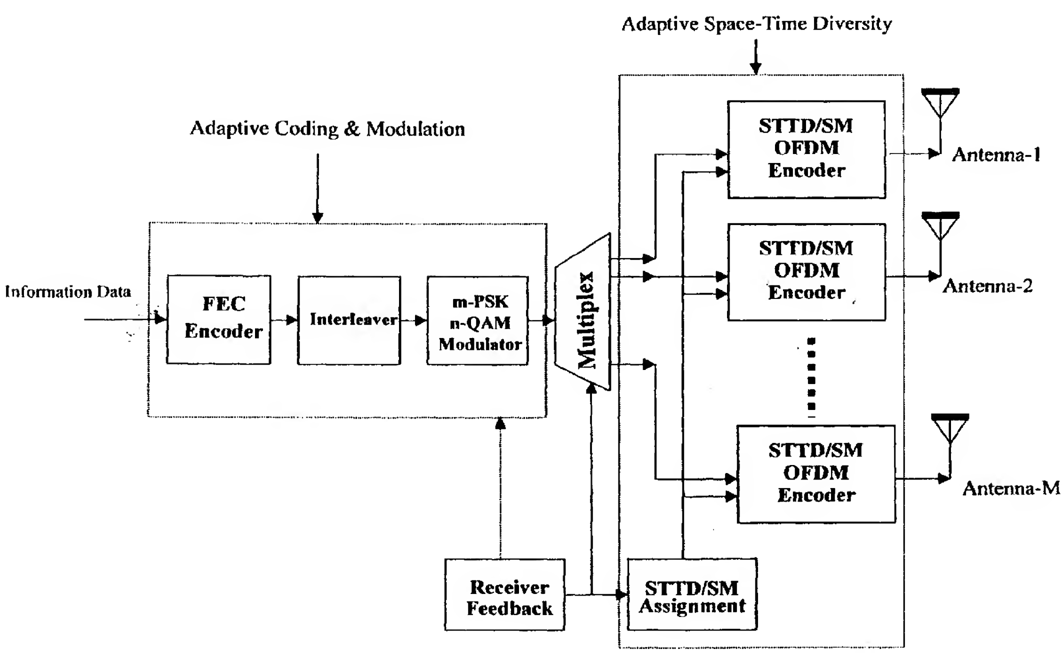

Figure-1

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Description

CROSS-REFERENCE TO RELATED PATENT APPLICATIONS

[0001] Reference is made copending patent application entitled: CHANNELS ESTIMATION FOR MULTIPLE INPUT - MULTIPLE OUTPUT, ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING (OFDM) SYSTEM, and whose contents are incorporated by reference with respect to channels estimation. This application is a utility patent application based on provisional patent application serial 30/229972, filed September 1, 2000.

BACKGROUND OF THE INVENTION

Field of the Invention

[0002] The invention relates to adapting time diversity and spatial diversity for use in an orthogonal frequency-division multiplexing (OFDM) environment, using a multiple input and multiple output (MIMO) structure.

Discussion of Related Art

[0003] A multiple input, multiple output (MIMO) structure has multiple communication channels that are used between transmitters and receivers. A space time transmitter diversity (STTD) system may be used on a MIMO structure, but it will not increase the data throughput. Indeed, for a high level configuration, the data rate may even reduce. In an STTD system, the transmitters deliver the same information content within consecutive symbol duration so that time diversity may be exploited. To efficiently use the multiple transmitters of the MIMO structure, however, the transmission data rate needs to be increased.

[0004] The most straightforward solution to increase the transmission data rate is to in forward error correction (FEC) dump independent data to each transmitter. A forward error correction (FEC) encoder produces in-phase and quadrature-phase data streams for the digital QAM modulator in accordance with a predetermined QAM constellation. The QAM modulator may perform baseband filtering, digital interpolation and quadrature amplitude modulation. The output of the QAM modulator is a digital intermediate frequency signal. A digital to analog (D/A) converter transforms the digital IF signal to analog for transmission.

[0005] The problem arises, however, as to how to safely recover the transmitted data. For a 2x2 system (two transmitters, two receivers) for example, after the channel information is obtained, the recovery process entails formulating two equations with two unknowns that need to be solved. The two unknowns may be determined only if the 2x2 channel is invertible. In practice, however, two situations may be encountered, i.e., the channel matrix is rank deficient so the unknowns cannot be determined or the frequency response channel matrix is invertible but has a very small eigen value.

[0006] The first situation arises when the channels are highly correlated, which may be caused either by not enough separation of the transmitters or by homology of the surroundings. For the second situation, although the equations are solvable, the solution can cause a high bit error rate (BER), because a scale up of the noise can result in an incorrect constellation point.

[0007] Orthogonal frequency-domain multiplexing (OFDM) systems were designed conventionally for either time diversity or for space diversity, but not both. The former will provide a robust system that combats signal fading but cannot increase the data rate capacity, while the latter can increase the data rate capacity but loses the system robustness. An OFDM signal contains OFDM symbols, which are constituted by a set of sub-carriers and transmitted for a fixed duration.

[0008] The MIMO structure may be used for carrying out time diversity for an OFDM system. For instance, when one transmitter transmits an OFDM signal, another transmitter will transmit a fully correlated OFDM signal to that transmitted by the one transmitter. The same OFDM signal is transmitted with, for instance, a fixed OFDM duration.

[0009] On the other hand, spatial diversity entails transmitting independent signals from different transmitters. Thus, transmitting two independent OFDM signals from two transmitters, respectively, results in a double data rate capacity from the parallel transmission that occurs.

[0010] When the signal to noise ratio (SNR) is low, the frame error rate (FER) is large, so that a data packet transmission will be decoded incorrectly and will need to be retransmitted. The quality of service (QoS) defines the number of times that the same packet can be retransmitted, eg., within an OFDM architecture. The OFDM system on a MIMO structure, therefore, should be adaptable to ensure that the QoS is maintained.

[0011] For any given modulation and code rate, the SNR must exceed a certain threshold to ensure that a data packet will be decoded correctly. When the SNR is less than that certain threshold, the bit error rate (BER) will be larger, which results in a larger FER. The larger the FER, the more retransmissions of the same packet will be required until the packet is decoded correctly. Thus, steps may need to be taken to provide the OFDM system with a higher

gain. If the SNR is at or above the threshold, then there is no need to increase the gain of the architecture to decode the data packets correctly. One challenge is to adapt the OFDM system to use time diversity when signal fading is detected as problematic and to use spatial diversity at other time to increase the data rate transfer.

[0012] In a conventional OFDM system, there are many OFDM modes, for examples are the 1k mode (1024 tones) and the half k mode(512 tones). For 1k mode, the number of sub-carriers is 1024 and for the half k mode, the number of sub-carriers is 512. The 1k mode is suitable for a channel with long delay and slow temporal fading, while the 512 mode is suitable for the channel with a short delay and fast temporal fading. But which mode will be used is really depending on the real environment.

[0013] A transaction unit of a conventional OFDM signal is an OFDM frame that lasts 10 ms. Each OFDM frame consists of 8 OFDM slots and each slot lasts 1.25 ms. Each OFDM slot consists of 8 OFDM symbols and some of the OFDM symbols will be the known preambles for access and channels estimation purposes. An OFDM super frame is made up of 8 OFDM frames and lasts 80 ms.

[0014] In addition to transmitted data, an OFDM frame contains a preamble, continual pilot sub-carriers, and transmission parameter sub-carriers/scattered sub-carriers. The preamble contains OFDM symbols that all used for training to realize timing, frequency and sampling clock synchronization acquisitions, channel estimation and a C/I calculation for different access points.

[0015] The continual pilot sub-carriers contain training symbols that are constant for all OFDM symbols. They are used for tracking the remaining frequency/sampling clock offset after the initial training.

[0016] The transmission parameter sub-carriers/scattered sub-carriers are dedicated in each OFDM symbol and reserved for signaling of transmission parameters, which are related to the transmission scheme, such as channel coding, modulation, guarding interval and power control. The transmission parameter sub-carriers are well protected and therefore can be used as scattered pilot sub-carriers after decoding.

[0017] One application for determining whether sub-carriers should be assigned to time diversity or spatial diversity is to conform with statistical analysis of traffic demands during particular times of the day, such as peak and off-peak. The OFDM system may preferably bias toward either time diversity or spatial diversity based on such a statistical analysis.

BRIEF SUMMARY OF THE INVENTION

[0018] One aspect of the invention pertains to employing adaptive STDD and spatial multiplexing (SM) based on comparing the channel condition of each sub-carrier with a threshold. When a sub-carrier is accommodated on channels that have a "well conditioned" channel matrix, spatial multiplexing may be used to create independent transmission paths and therefore increase the data rate. A "well conditioned" channel matrix arises when the smallest eigen value is not too small as compared to a threshold value, such as the noise power increase when multiplied by its inverse. For those sub-carriers whose channel matrices have smaller eigen values, the receiver cannot recover the parallel transmitted information symbols. As a result, STTD is used to guarantee a robust system.

[0019] Encoders associated with the transmitter side encode or classify sub-carriers in accordance with one of two groups based on a feedback signal; one of the groups is to forward error correction (FEC) time diversity and the other of the two groups is to forward error correction (FEC) spatial diversity. This grouping is based on results from a comparison made at the receiver side between a threshold value and either a calculated smallest eigen value of a frequency response matrix, the smallest element in a diagonal of the matrix, or a ratio of the largest and smallest eigen values in the matrix.

[0020] The threshold value is based on the transmitter and receiver antenna configuration, environmental constraints of the OFDM communication system, and/or on statistical analysis of communication traffic demands. The estimate value is derived from channel estimation of multiple channels of multi-input multi-output (MIMO) type systems.

[0021] Time diversity is used to reduce adverse signal fading. Spatial diversity is used to increase the data rate, which time diversity cannot do. When sub-carriers use time diversity, it means that signal fading is strong so that parallel transmission of data packets can not be done to overcome the insufficient gain problem. Instead, time diversity is used to get the necessary gain for the OFDM system, even though the data rate capacity suffers. An SNR gain is assured with time diversity, because of the orthogonality matrix pattern inherent among transmitted samples in the OFDM system. On the other hand, when sub-carriers use spatial diversity, signal fading is weak so that parallel transmissions may occur to increase the data rate capacity. Thus, there is no need to increase the gain of the OFDM system, which means that the data rate may be increased.

[0022] In operation, the OFDM system of the invention may start transmission of data packets with either time diversity or spatial diversity. The receiver side will estimate the channels and decode the data packets. After the channel information, is obtained, the receiver side will calculate the eigen values of the channel matrices to the extent possible. The controller then determines whether the sub-carrier to use time diversity or spatial diversity based on one of three criteria (only one of which is dependent upon the eigen value calculation). The receiver then reports back or feedbacks to the

transmitter side with this information, i.e., about whether the sub-carrier is to use time diversity or spatial diversity so as to trigger the next round of transmission accordingly.

BRIEF DESCRIPTION OF THE SEVERAL VIEWS OF THE DRAWING

[0023] For a better understanding of the present invention, reference is made to the following description and accompanying drawings, while the scope of the invention is set forth in the appended claims.

[0024] Fig. 1 is a schematic representation of a generic multi-input, multi-output orthogonal frequency-division multiplexing transmitter in accordance with an embodiment of the invention.

[0025] Fig. 2 is a schematic representation of an orthogonal frequency-division multiplexing symbol.

[0026] Fig. 3 is a space time transmitter diversity (STTD) orthogonal frequency-division multiplexing (OFDM) encoder for loading data to a sub-carrier in G1 which will be specified in the forthcoming sections.

[0027] Fig. 4 is a spatial multiplexing (SM) orthogonal frequency-division multiplexing (OFDM) encoder for loading data to a sub-carrier in G2 which will be specified in the forthcoming sections.

[0028] Fig. 5 is a schematic representation of two pure STTD transmitters that save one half of the IFFT computation.

[0029] Fig. 6 is a schematic representation of four pure STTD transmitters that save three fourths of the IFFT computation.

[0030] Fig. 7 is a schematic representation a generic receiver structure.

[0031] Fig. 8 is a schematic representation of configurations of a two receiver antenna case and a three receiver antenna case.

DETAILED DESCRIPTION OF THE INVENTION

[0032] The invention concerns a practical time and spatial diversity combination that fits into an OFDM system. The OFDM system of the invention can automatically adapt the channel variation and make trade off between time diversity and spatial diversity. In an exemplary environment, the data rate can be increased 1.8 times for 2x2 configuration (2 transmitters, 2 receivers), which gives 80 Mbps, and 2.7 times for 3x3 configuration)3 transmitters, 3 receivers) which gives 121 Mbps within 6MHz, while keep the robustness of the system.

[0033] Turning to the drawing, Figure 1 shows a generic MIMO and OFDM transmitter system. In the figure, STTD and SM are the abbreviations of Space-Time-Transmitter Diversity and Spatial Multiplexing. The MIMO OFDM is configured as two level adaptations as shown in Figure 1, namely, space/time diversity adaptation and coding/modulation adaptation. The space/time diversity adaptation is determined by the carrier to interference power ratio or signal to noise power ratio.

[0034] Information data is fed into adaptive coding modulation, the modulation is multiplexed and fed into adaptive space/time diversity encoding and assignment. A receiver feedback to provide feedback signals to the adaptive coding of modulation, multiplexer and adaptive space/time diversity is also provided. The multiplexed signals in the adaptive space/time diversity pass through STTD/SM OFDM encoders and the encoded signals transmit to associated antennas. The adaptive coding and modulation includes a forward error correction (FEC) encoder, an interleaver and an m-PSK modular.

[0035] If x MHz bandwidth is available, then Orthogonal Frequency Division Multiplexing OFDM is to chop this whole spectrum into many small pieces of equal width and each of them will be used as a carrier. The width of the piece will be determined by delay spread of the targeted environment.

[0036] The STTD/OFDM encoder is responsible for the assignment of the constellation points to each sub-carrier. For M transmitters, M OFDM symbols data are loaded in general (so the bit loading will be calculated according to this number), but it will depend on the STTD structure. Figure 2 illustrates one OFDM symbol.

[0037] For each sub-carrier that is indexed k, its loading will be determined by its corresponding channel condition. For N receivers, the frequency channel responses may be represented by an MxN matrix, say H(k). The channel condition will be described by one of the following 3 criteria.

1. Smallest eigen value of $H(k)H(k)^*$
2. Smallest element of the diagonal of $H(k)H(k)^*$
3. The ratio of largest and smallest eigen values of $H(k)H(k)^*$

[0038] A set of thresholds for each criterion and for each system configuration is used. These thresholds will be service parameters and can be used as quality of service (QoS) or billing purposes.

[0039] With each criterion and a given threshold, all the sub-carriers will be classified into two groups G1 and G2 by a controller at the receiver side. The controller directs the transmission of a feedback signal indicative of the result of the classification. The feedback signal is received at the transmitter side and interpreted by a controller at the transmitter

side. The sub-carriers classified in G1 will use STTD encoder at the transmission side while those classified in G2 will use the SM encoder at the transmission side.

[0040] After the subcarriers have been classified into the two groups G1 and G2, the modulation scheme on each sub-carrier will be determined by the estimated C/I (carrier to interference ratio) or SNR (signal to noise ratio). As a result, a modulation scheme, such as of QPSK or m-PSK or various QAM, will be selected to satisfy QoS (quality of service) based on the determination made by the estimated C/I or SNR. This is another level adaptation that may maximize the throughput gain.

[0041] For instance, when the QoS is defined, the FER (frame error rate) may be ten percent. The goal is to choose a modulation scheme according to the perceived C/I or SNR to satisfy this QoS, yet still maximizing the throughput of data flow. To achieve this, a pre-defined look-up table may be accessed that is in accordance with various QoS.

[0042] In determining which modulation scheme will satisfy the criteria, the C/I or SNR estimation is done during mobile access, after looking for the strongest signal from the base station first. Based on such knowledge and estimation, one is able to get a rough idea as to which modulation scheme should be used. Regardless of the modulation scheme selected initially, the invention is configured to automatically adapt toward whichever modulation scheme represents the optimal modulation.

[0043] Fig. 3 shows how to load data on sub-carrier k for a situation involving 2 transmitters for example. This data loading is done within a pair of OFDM symbols. As can be appreciated, apparently one sample has been transmitted twice within 2 OFDM symbols duration via 2 transmitters. Thus, the data rate is the same as for the one transmitter OFDM system.

[0044] Fig. 4 shows how to load data on sub-carrier k in G2 for a situation involving 2 transmitters. In this case, each transmitter transmits independent data and therefore the data rate is double for 2 transmitters and M times for M transmitters.

[0045] The adaptive time diversity and spatial diversity for OFDM works as follows. Starting out, an STTD mode is used for all sub-carriers. The receiver estimates the channel profiles and then directs a feedback of its preference either to STTD or spatial multiplexing (SM) on each sub-carrier.

[0046] The whole sub-carrier indices $\{K_{min}, K_{min}+1, ..., K_{max}\}$ are then divided into two disjoint subsets I_{sttd} and I_{sm}. The one with fewer elements will be the feedback to the transmitters. The extreme case is that one of them is an empty set, which means use of either pure STTD or pure SM. As in the pure STTD system, the transmitters always consider two OFDM symbols as the basic transmission unit for 2x2 configuration and M OFDM symbols for a system has M transmitters.

[0047] The number of input bits, however, needs to be calculated according to a modulation scheme and a dynamic distribution of I_{sttd} and I_{sm}. More precisely, the number of bits needed for the two consecutive OFDM symbols is $2x|I_{sttd}|L + 4x|I_{sm}|L$, where L is the modulation level which equals to 2, 3, 4 5, 6, 7, 8.

[0048] When a granularity problem arises, the two OFDM symbols are repacked to fit the granularity by removing some sub-carriers from I_{sm} into I_{sttd}. This may sacrifice the data rate somewhat, but keep the system robust.

[0049] In the receiver side, a quadrature amplitude modulation QAM de-mapping block is used to de-map the received data according to I_{sttd} and I_{sm}.

[0050] STTD is the baseline of the service quality. This means that when parallel transmission is carried out in the designated communication channels, then it is guaranteed parallel transmission, because the BER or FER will be controlled to achieve the necessary QoS. The transmitters will propagate the transmissions at the same constant power and the modulation will be the same for each transmitter. Thus, no power pouring technique needs to be employed.

[0051] Three thresholds are used to classify the sub-carriers. Indeed, the threshold can be used as a service parameter and tuned aggressive to either STTD mode or SM mode according to customer demand, i.e., based on statistical analysis of that demand.

[0052] As an example, for the case where the smallest eigen value is used as the threshold in a 2x2 configuration (2 transmitters, 2 receivers), there is a 60 % opportunity to do parallel transmission with 0.5 as the threshold value, which may be scale the noise 3 dB up. for a 2x4 configuration (2 transmitters, 4 receivers), there is an 80 % opportunity to do parallel transmission with 1 as the threshold value, which may even reduce the noise.

[0053] Fig. 5 shows a special, but very practical situation, which shows two pure STTD transmitters that save ½ of an inverse fast Fourier transform (IFFT) computation. The present invention may automatically switch to this scenario in a vulnerable environment involving 2 transmitters.

[0054] Conventionally, one would expect each transmitter to transmit 2 OFDM symbols every 2 OFDM symbol duration. Thus, there are 4 OFDM symbols transmitted for every 2 OFDM duration that go through a respective independent IFFT computation engine. This means that a complex number IFFT computation is expected to be conducted four times.

[0055] For a pure STTD implementation with 2 and 4 transmit antennas, the computational efficient implementation is shown in Figures 5 and 6 respectively. The scheme in Figure 5 requires ½ of the IFFT computation and the scheme in Figure 6 requires ¼ of the IFFT computation as compared with a straightforward implementation that performs the

computations separately.

[0056] In accordance with Fig. 5, however, there is data crossing between two transmitters, which saves two IFFT computations. Yet, it provides four IFFT outputs, which is exactly the same results where four independent IFFTs are used. Although four IFFT operations are shown in Fig. 5, they are operating on real vectors, which means the computational complexity of a real IFFT equals the complex IFFT with a half size. Therefore, the computational time saving comes from the relationship between IFFT on a vector and its conjugate.

[0057] In Fig. 5, the bits are coded bits, which are the input to variable M-PSK/QAM mapping.. The mapping will map the bits to the corresponding constellation points according to the Gray rule; constellation points here refer to any modulation scheme, such as QPSK, m-PSK, QAM, etc. The constellation vector will be inserted with a pilot into a multiplex and then into first in first out (FIFO) buffers.

[0058] The designations S_0, S_1, S_2, S_3, S_{2046}, S_{2047}, in the FIFO buffer represent complex vectors. The function Re {} refers to just taking the real part of the complex vector. The designation Im{} refers to just taking the imaginary part of the complex vector. The real and imaginary parts are fed as input into IFFTs. The designation D/A refers to a digital to analog converter.

[0059] The transmission order for the first transmitter is OFDM symbol b and then d ...; the transmission order for the second transmitter is OFDM symbol g and then f etc. Before each OFDM symbol is transmitted, the cyclic extension will be appended somewhere in the OFDM symbol.

[0060] Periodically inserted preambles will serve for the timing recovery, framing, frequency offset estimation , clock correction and overall channel estimation The estimated channel samples will be used for the continuous spectrum channel reconstruction. Pilot symbols will serve for phase correction. final tuning of channel estimation.

The mathematical equivalence for Fig. 5 is as follows

$$b = IFFT\begin{bmatrix} S_0 \\ S_2 \\ \vdots \\ S_{2046} \end{bmatrix}, \quad d = IFFT\begin{bmatrix} -S_1^* \\ -S_3^* \\ \vdots \\ -S_{2047}^* \end{bmatrix}, \quad f = IFFT\begin{bmatrix} S_1 \\ S_3 \\ \vdots \\ S_{2047} \end{bmatrix}, \quad g = IFFT\begin{bmatrix} S_0^* \\ S_2^* \\ \vdots \\ S_{2046}^* \end{bmatrix}$$

[0061] Fig. 6 shows four Pure STTD Transmitters that represents a rate 3/4 STTD encoder as:

Tx1	S(0)	-S(1)*	S(2)*/sqrt(2)	S(2)/sqrt(2)
Tx2	S(1)	S(0)*	S(2)*/sqrt(2)	- S(2)/sqrt(2)
Tx3	S(2)/sqrt(2)	S(2)/sqrt(2)	-Re{S(0)}+jIm{S(1)}	-Re{S(1)}+jIm{S(0)
Tx4	S(2)/sqrt(2)	S(2)/sqrt(2)	Re(S(1)}+jIm(S(0)}	-Re{S(0)}-jIm{S(1)
Time	[0 T]	[T 2T]	[2T 3T]	[3T 4T]

Such an STTD encoder encodes every 3 OFDM symbols into 4 OFDM symbols and transmits to 4 antennas. Figure 6 scheme requires 1/4 IFFT computation compared to the straightforward implementation. The reason why computation is saved is for the same reasons as in Fig. 5. The parameters there are defined respectively as follows:

$$b = IFFT\begin{bmatrix} S_0 \\ S_3 \\ \vdots \\ S_{3069} \end{bmatrix}, \quad g = IFFT\begin{bmatrix} S_0^* \\ S_3^* \\ \vdots \\ S_{3069}^* \end{bmatrix}$$

$$f = IFFT\begin{bmatrix} S_1 \\ S_4 \\ \vdots \\ S_{3070} \end{bmatrix}, d = IFFT\begin{bmatrix} S_1^* \\ S_4^* \\ \vdots \\ S_{3070}^* \end{bmatrix}$$

$$q = IFFT\begin{bmatrix} S_2 \\ S_5 \\ \vdots \\ S_{3071} \end{bmatrix}, u = IFFT\begin{bmatrix} S_2^* \\ S_5^* \\ \vdots \\ S_{3071}^* \end{bmatrix}$$

[0062] Fig. 7 is an abstract diagram of a generic receiver structure.
STTD/SM OFDM decoder is sub-carrier based decoder. The structure and configuration of the STTD/SM OFDM decoder will depend on the architecture configuration.
Suppose sub-carrier m is STTD coded, i.e. m belongs to G1.

[0063] For a 2x2 configuration:
S(2m) and S(2m+1) are decoded by solving the following equations

$$\begin{bmatrix} y_1(q,m) \\ y_1(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{11}(q,m) & h_{21}(q,m) \\ h_{21}(q,m)^* & -h_{11}(q,m)^* \end{bmatrix}\begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_1(q,m) \\ n_1(q+1,m) \end{bmatrix}$$

[0064] The assumption here is that the even indexed sample S(2m) is transmitted in qth OFDM and the odd indexed sample S(2m+1) is transmitted in (q+1)th OFDM symbol.

[0065] There are 4 equations and two unknowns. So a least mean square solution can be obtained by multiplying the coefficient matrix to the received data vector. With the above two pairs, we will get two estimated of the same pair of samples. Their average will be the output of the decoder.

[0066] More statistics are performed after regrouping the equations. In fact, every pair of the equations will result a solution, every 3 equations also provide a new estimation, and all the equations will give a solution too. There are 10 combinations in total and therefore 10 estimation with these 4 equations. Their average or partial average will be used as the solution.

[0067] A 2x3 configuration is similar to 2x2, involving 6 equations:

$$\begin{bmatrix} y_1(q,m) \\ y_1(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{11}(q,m) & h_{21}(q,m) \\ h_{21}(q,m)^* & -h_{11}(q,m)^* \end{bmatrix}\begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_1(q,m) \\ n_1(q+1,m) \end{bmatrix}$$

$$\begin{bmatrix} y_2(q,m) \\ y_2(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{12}(q,m) & h_{22}(q,m) \\ h_{22}(q,m)^* & -h_{12}(q,m)^* \end{bmatrix}\begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_2(q,m) \\ n_2(q+1,m) \end{bmatrix}$$

$$\begin{bmatrix} y_3(q,m) \\ y_3(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{13}(q,m) & h_{23}(q,m) \\ h_{23}(q,m)^* & -h_{13}(q,m)^* \end{bmatrix}\begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_3(q,m) \\ n_3(q+1,m) \end{bmatrix}$$

[0068] For a 2x4 configuration, there are 8 equations:

$$\begin{bmatrix} y_1(q,m) \\ y_1(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{11}(q,m) & h_{21}(q,m) \\ h_{21}(q,m)^* & -h_{11}(q,m)^* \end{bmatrix} \begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_1(q,m) \\ n_1(q+1,m) \end{bmatrix}$$

$$\begin{bmatrix} y_2(q,m) \\ y_2(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{12}(q,m) & h_{22}(q,m) \\ h_{22}(q,m)^* & -h_{12}(q,m)^* \end{bmatrix} \begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} z_2(q,m) \\ z_2(q+1,m) \end{bmatrix}$$

$$\begin{bmatrix} y_3(q,m) \\ y_3(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{13}(q,m) & h_{23}(q,m) \\ h_{23}(q,m)^* & -h_{13}(q,m)^* \end{bmatrix} \begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_3(q,m) \\ n_3(q+1,m) \end{bmatrix}$$

$$\begin{bmatrix} y_4(q,m) \\ y_4(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{14}(q,m) & h_{24}(q,m) \\ h_{24}(q,m)^* & -h_{14}(q,m)^* \end{bmatrix} \begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_4(q,m) \\ n_4(q+1,m) \end{bmatrix}$$

[0069] For a 4x2 configuration, there are 8 equations and 3 unknowns

$$\begin{bmatrix} y_1(q,m) \\ y_1(q+1,m) \\ y_1(q+2,m) \\ y_1(q+3,m) \end{bmatrix} = \begin{bmatrix} s(3m-3)) & s(3m-2) & \frac{s(3m-1)}{\sqrt{2}} & \frac{s(3m-1)}{\sqrt{2}} \\ -s(3m-2)* & s(3m-3)* & \frac{s(3m-1)}{\sqrt{2}} & -\frac{s(3m-1)}{\sqrt{2}} \\ \frac{s(3m-1)*}{\sqrt{2}} & \frac{s(3m-1)*}{\sqrt{2}} & \eta(m) & \kappa(m) \\ \frac{s(3m-1)*}{\sqrt{2}} & -\frac{s(3m-1)*}{\sqrt{2}} & \nu(m) & \zeta(m) \end{bmatrix} \begin{bmatrix} h_{11}(m) \\ h_{21}(m) \\ h_{31}(m) \\ h_{41}(m) \end{bmatrix} + \begin{bmatrix} n_{11} \\ n_{21} \\ n_{31} \\ n_{41} \end{bmatrix}$$

where

$\eta(m)$ = -Re(s(3(m - 1))) + j Im(s(3(m - 1) + 1)),
$\kappa(m)$ = -Re(s(3(m - 1) + 1)) + j Im ag(s(3(m - 1))),
$\nu(m)$ = Re(s(3(m - 1) + 1)) + j Im(s(3(m - 1))),
$\zeta(m)$ = -Re(s(3(m - 1))) - j Im ag(s(3(m - 1) + 1)),

h_{k1} (m) is the frequency channel response of the channel between transmitter k and receiver l. Similarly, the received data for the 4x2 configuration is

$$\begin{bmatrix} y_2(q,m) \\ y_2(q+1,m) \\ y_2(q+2,m) \\ y_2(q+3,m) \end{bmatrix} = \begin{bmatrix} s(3m-3)) & s(3m-2) & \frac{s(3m-1)}{\sqrt{2}} & \frac{s(3m-1)}{\sqrt{2}} \\ -s(3m-2)* & s(3m-3)* & \frac{s(3m-1)}{\sqrt{2}} & -\frac{s(3m-1)}{\sqrt{2}} \\ \frac{s(3m-1)*}{\sqrt{2}} & \frac{s(3m-1)*}{\sqrt{2}} & \eta(m) & \kappa(m) \\ \frac{s(3m-1)*}{\sqrt{2}} & -\frac{s(3m-1)*}{\sqrt{2}} & \nu(m) & \zeta(m) \end{bmatrix} \begin{bmatrix} h_{12}(m) \\ h_{22}(m) \\ h_{32}(m) \\ h_{42}(m) \end{bmatrix} + \begin{bmatrix} n_{11} \\ n_{21} \\ n_{31} \\ n_{41} \end{bmatrix}$$

[0070] The solution will be the least mean square solution by enumerating all possibilities. Suppose instead that sub-carrier m is SM Coded, i.e. m belongs to G2. For a 2x2 configuration, there are 4 equations and 4 unknowns:

$$\begin{bmatrix} y_1(q,m) \\ y_2(q,m) \end{bmatrix} = \begin{bmatrix} h_{11}(q,m) & h_{21}(q,m) \\ h_{12}(q,m) & h_{22}(q,m) \end{bmatrix} \begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_1(q,m) \\ n_1(q,m) \end{bmatrix}$$

$$\begin{bmatrix} y_1(q+1,m) \\ y_2(q+1,m)^* \end{bmatrix} = \begin{bmatrix} h_{11}(q,m) & h_{21}(q,m) \\ h_{12}(q,m) & h_{22}(q,m) \end{bmatrix} \begin{bmatrix} s(2m+2) \\ s(2m+3) \end{bmatrix} + \begin{bmatrix} n_2(q+1,m) \\ n_2(q+1,m) \end{bmatrix}$$

[0071] So the 4 unknowns can be estimated by the least mean square solutions.
[0072] For a 2x3 configuration, there are 6 equations and 4 unknowns.

$$\begin{bmatrix} y_1(q,m) \\ y_2(q,m) \\ y_3(q,m) \end{bmatrix} = \begin{bmatrix} h_{11}(q,m) & h_{21}(q,m) \\ h_{12}(q,m) & h_{22}(q,m) \\ h_{13}(q,m) & h_{23}(q,m) \end{bmatrix} \begin{bmatrix} s(2m) \\ s(2m+1) \end{bmatrix} + \begin{bmatrix} n_1(q,m) \\ n_2(q,m) \\ n_3(q,m) \end{bmatrix}$$

$$\begin{bmatrix} y_1(q+1,m) \\ y_2(q+1,m) \\ y_3(q+1,m) \end{bmatrix} = \begin{bmatrix} h_{11}(q+1,m) & h_{21}(q+1,m) \\ h_{12}(q+1,m) & h_{22}(q+1,m) \\ h_{13}(q+1,m) & h_{23}(q+1,m) \end{bmatrix} \begin{bmatrix} s(2m+2) \\ s(2m+3) \end{bmatrix} + \begin{bmatrix} n_1(q+1,m) \\ n_2(q+1,m) \\ n_3(q+1,m) \end{bmatrix}$$

[0073] For a 2x4 configuration, there are 8 equations and 4 unknowns
[0074] For a 3x3 configuration, there are 9 equations and 9 unknowns.
[0075] In accordance with the inventive architecture, the data rate can be as high as 70 Mbps for 2x2 and 120 Mbps for 3x3 within 6 MHz spectrum.
[0076] An exemplary optimal threshold value for a 2x2 configuration is 0.5. An exemplary optimal threshold value for a 2x4 configuration is 1.0. An exemplary optimal threshold value for a 3x3 configuration is 1.2. An exemplary optimal threshold value for a 2x3 configuration is 1.0. By exemplary optimal threshold value, the intent is to attain a value that has a trade-off between time and spatial diversity that yields both a relatively high robustness and relatively high data packet rate transfer.
[0077] As can be appreciated for each of the afore-mentioned configurations, there are a certain number of equations and a certain number of unknowns. In an over-determined system, the number of equations is greater than the number of unknowns. Thus, for a 2x2 configuration, there are two unknowns but four equations may be formulated. If there is no noise, any two of them (six pairs), or any three of them (four triples) or all of the four equations (one quadratic) will give the same answer. The difference is when noise is present, because the combinations with then give different solutions. Since some of the solutions may be good while others are bad, different combinations are chosen, but those combinations that result in large derivations are to be avoided. The idea is to use a sub-set of the over determined linear equations to estimate the solution and then average all the possible solutions that seem viable. The averaging may be done with a least mean square solution, which is a conventional mathematical technique.
[0078] Figure 8 compares a two receiver antenna case and a three receiver antenna case. With respect to the three receiver antenna case, the number of receiver antennas is greater than the number of transmitter antennas. As a consequence, the receiver has additional redundancy, the receiver has various configurations, and the configurations yield several different decoding results. The most reliable solution can be selected from among them or all the solutions may be averaged to obtain a final result.
[0079] While the foregoing description and drawings represent the preferred embodiments of the present invention, it will be understood that various changes and modifications may be made without departing from the spirit and scope

of the present invention.

Claims

1. An apparatus for use with an adaptive orthogonal frequency division-multiplexing (OFDM) system that uses multiple input multiple output (MIMO) structure to transmit OFDM signals from a plurality of transmitters to a plurality of receivers, the OFDM signal having an OFDM frame of a duration, the OFDM frame having data packets and a plurality of OFDM slots, each of the OFDM slots having a plurality of OFDM symbols that include a plurality of sub-carriers, the apparatus comprising:

   a receiver that responds to receipt of the OFDM signal by making a determination as to whether time diversity or spatial diversity should be used for subsequent transmissions and transmits a feedback signal indicative of that determination, an implementation of the time diversity resulting in a better robustness to counter signal fading than if the spatial diversity were implemented and an implementation of spatial diversity resulting in an increase in a rate of data packet transfer over that if the time diversity were implemented, because the OFDM signals that are transmitted over multiple ones of the transmitters are independent of each other for the spatial diversity and correspond to each other for the time diversity.

2. An apparatus as in claim 1, wherein the receiver includes a controller that makes the determination based on a comparison of a channel condition with a threshold, the channel condition being based on a frequency response channel matrix that is derived from OFDM symbols.

3. An apparatus as in claim 2, wherein the channel condition is based on a calculation of a smallest eigen value of the frequency response channel matrix.

4. An apparatus as in claim 2, wherein the channel condition is based on a determination of a smallest element in a diagonal of the frequency response channel matrix.

5. An apparatus as in claim 2, wherein the channel condition represents a ratio of largest and smallest eigen values of the channel matrix.

6. An apparatus as in claim 2, wherein the channel condition is based on one of three criteria selected from a group consisting of a calculation of smallest eigen values of the channel matrix, a smallest element in a diagonal of the channel matrix, and a ratio of largest and smallest eigen values of the channel matrix.

7. An apparatus as in claim 2, further comprising a channel estimator that forms the frequency response channel matrix.

8. An apparatus as in claim 2, wherein the controller is configured to classify the sub-carriers into one of two groups in accordance with the channel condition, one of the two groups being indicative of time diversity and the other of the two groups being indicative of spatial diversity, the controller being further configured to determine a modulation scheme on each of the classified sub-carriers based on an estimated ratio selected from a further group consisting of a carrier to interference ratio and a signal to noise ratio.

9. An apparatus for use with an adaptive orthogonal frequency division-multiplexing (OFDM) system that uses multiple input multiple output (MIMO) structure to transmit OFDM signals from a plurality of transmitters to a plurality of receivers, the OFDM signal having an OFDM frame of a duration, the OFDM frame having data packets and a plurality of OFDM slots, each of the OFDM slots having a plurality of OFDM symbols that include a plurality of sub-carriers, the apparatus comprising

   at least one controller configured and arranged to respond to a feedback signal to direct an encoder to assign constellation points to the sub-carriers in accordance with a channel condition so as to classify each of the sub-carriers into one of two groups, the encoder including a space time transmitter diversity (STTD) encoder and a spatial multiplexing (SM) encoder, the STTD encoder being arranged to encode the sub-carriers classified in one of the groups in accordance with time diversity and the SM encoder being arranged to encode the sub-carriers classified in the other of the groups in accordance with spatial diversity, an implementation of the time diversity resulting in a better robustness to counter signal fading than if the spatial diversity were implemented and an implementation of spatial diversity resulting in an increase in a rate of data packet transfer over that if the time diversity were implemented, because the OFDM signals that are transmitted over multiple ones of the transmitters

are independent of each other for the spatial diversity and correspond to each other for the time diversity.

10. An apparatus as in claim 9, wherein the controller is configured to determine a modulation scheme on each of the sub-carriers based on an estimated ratio selected from a further group consisting of a carrier to interference ratio and a signal to noise ratio.

11. An apparatus for use with an adaptive orthogonal frequency division multiplexing (OFDM) system that uses multiple input multiple output (MIMO) structure to transmit OFDM signals from a plurality of transmitters to a plurality of receivers, the OFDM signal having an OFDM frame of a duration, the OFDM frame having data packets and a plurality of OFDM slots, each of the OFDM slots having a plurality of OFDM symbols that include a plurality of sub-carriers, the apparatus comprising:

controllers configured and arranged to direct transmission and reception in accordance with OFDM, the controllers including those associated with the reception that are configured to responds receipt of the OFDM signal by making a determination as to whether time diversity or spatial diversity should be used for subsequent transmissions and transmits a feedback signal indicative of that determination, an implementation of the time diversity resulting in a better robustness to counter signal fading than if the spatial diversity were implemented and an implementation of spatial diversity resulting in an increase in a rate of data packet transfer over that if the time diversity were implemented, because the OFDM signals that are transmitted over multiple ones of the transmitters are independent of each other for the spatial diversity and correspond to each other for the time diversity, the controllers associated with the reception being configured to direct that transmission of at least one feedback signal occur that reflects the determination, the controllers including those associated with the transmission that are responsive to receipt of the feedback signal to direct an encoder to assign constellation points to the sub-carriers in accordance with a channel condition so as to classify each of the sub-carriers into one of two groups, the encoder including a space time transmitter diversity (STTD) encoder and a spatial multiplexing (SM) encoder, the STTD encoder being arranged to encode the sub-carriers classified in one of the groups in accordance with the time diversity and the SM encoder being arranged to encode the sub-carriers classified in the other of the groups in accordance with the spatial diversity.

12. An apparatus as in claim 11, wherein the controller is configured to determine a modulation scheme on each of the sub-carriers based on an estimated ratio selected from a further group consisting of a carrier to interference ratio and a signal to noise ratio.

13. An apparatus as in claim 12, wherein the controllers associated with the reception are configured to make a calculation of eigen values of channel matrices to make a determination as to which sub-carriers are to use the time diversity to reduce signal fading forward error correction (FEC) during a subsequent transmission and which sub-carriers are to use the spatial diversity to increase a rate of data transfer during the subsequent transmission, the controllers associated with the reception being configured to make the determination based on a comparison between a threshold and at least one of three criteria and to direct transmission of a feed back signal indicative of a result of the determination, at least one of the criteria being based on the calculation, at least another of the criteria being based on elements of a diagonal of at least one of the channel matrices.

14. An apparatus as in claim 12, wherein the controllers associated with the reception are configured so to make the determination based on a comparison of a channel condition with a threshold, the channel condition being based on a frequency response channel matrix that is derived from OFDM symbols.

15. An apparatus as in claim 14, wherein the channel condition represents a calculation of a smallest eigen value of the frequency response channel matrix.

16. An apparatus as in claim 14, wherein the channel condition represents a determination of a smallest element in a diagonal of the frequency response channel matrix.

17. An apparatus as in claim 14, wherein the channel condition represents a ratio of largest and smallest eigen values of the channel matrix.

18. An apparatus as in claim 14, wherein the channel condition represents one of three criteria selected from a group consisting of a calculation of smallest eigen values of the channel matrix, a smallest element in a diagonal of the channel matrix, and a ratio of largest and smallest eigen values of the channel matrix.

19. An apparatus as in claim 14, further comprising a channel estimator that forms the frequency response channel matrix.

20. A method for use with an adaptive orthogonal frequency division -multiplexing (OFDM) system that uses multiple input multiple output (MIMO) structure to transmit OFDM signals from a plurality of transmitters to a plurality of receivers, the OFDM signal having an OFDM frame of a duration, the OFDM frame having data packets and a plurality of OFDM slots, each of the OFDM slots having a plurality of OFDM symbols that include a plurality of sub-carriers, the method comprising:

responding to receipt of the OFDM signal by making a determination as to whether time diversity or spatial diversity should be used for subsequent transmissions and transmits a feedback signal indicative of that determination, an implementation of the time diversity resulting in a better robustness to counter signal fading than if the spatial diversity were implemented and an implementation of spatial diversity resulting in an increase in a rate of data packet transfer over that if the time diversity were implemented, because the OFDM signals that are transmitted over multiple ones of the transmitters are independent of each other for the spatial diversity and correspond to each other for the time diversity.

21. A method as in claim 20, further comprising making the determination based on a comparison of a channel condition with a threshold, the channel condition being based on a frequency response channel matrix that is derived from OFDM symbols.

22. A method as in claim 21, further comprising calculating a smallest eigen value of the frequency response channel matrix basing the channel condition on the calculating.

23. An method as in claim 21, further comprising determining a smallest element in a diagonal of the frequency response channel matrix and basing the channel condition on the determining.

24. A method as in claim 21, further comprising calculating a ratio of largest and smallest eigen values of the channel matrix and basing the channel condition on the ratio.

25. A method as in claim 21, further comprising basing the channel condition on one of three criteria selected from a group consisting of a calculation of smallest eigen values of the channel matrix, a smallest element in a diagonal of the channel matrix, and a ratio of largest and smallest eigen values of the channel matrix.

26. A method as in claim 20, further comprising classifying the sub-carriers into two groups one of the two groups being indicative of time diversity and the other of the two groups being indicative of spatial diversity. determining a modulation scheme on each of the classified sub-carriers based on an estimated ratio selected from a further group consisting of carrier to interference ratio and signal to noise ratio.

27. A method for use with an adaptive orthogonal frequency division-multiplexing (OFDM) system that uses multiple input multiple output (MIMO) structure to transmit OFDM signals from a plurality of transmitters to a plurality of receivers, the OFDM signal having an OFDM frame of a duration, the OFDM frame having data packets and a plurality of OFDM slots, each of the OFDM slots having a plurality of OFDM symbols that include a plurality of sub-carriers, the method comprising

responding to a feedback signal to direct an encoder to assign constellation points to the sub-carriers in accordance with a channel condition so as to classify each of the sub-carriers into one of two groups, the encoder including a space time transmitter diversity (STTD) encoder and a spatial multiplexing (SM) encoder, the STTD encoder being arranged to encode the sub-carriers classified in one of the groups in accordance with time diversity and the SM encoder being arranged to encode the sub-carriers classified in the other of the groups in accordance with spatial diversity, an implementation of the time diversity resulting in a better robustness to counter signal fading than if the spatial diversity were implemented and an implementation of spatial diversity resulting in an increase in a rate of data packet transfer over that if the time diversity were implemented, because the OFDM signals that are transmitted over multiple ones of the transmitters are independent of each other for the spatial diversity and correspond to each other for the time diversity.

28. A method as in claim 27, further comprising classifying the sub-carriers into two groups, one of the two groups being indicative of time diversity and the other of the two groups being indicative of spatial diversity, determining a modulation scheme on each of the classified sub-carriers based on an estimated ratio selected from a further group consisting of a carrier to interference ratio and a signal to noise ratio.

29. A method for use with an adaptive orthogonal frequency division-multiplexing (OFDM) system that uses multiple input multiple output (MIMO) structure to transmit OFDM signals from a plurality of transmitters to a plurality of receivers, the OFDM signal having an OFDM frame of a duration, the OFDM frame having data packets and a plurality of OFDM slots, each of the OFDM slots having a plurality of OFDM symbols that include a plurality of sub-carriers, the method comprising:

directing transmission and reception in accordance with OFDM by using controllers, the controllers including those associated with the reception responding to receipt of the OFDM signal by making a determination as to whether time diversity or spatial diversity should be used for subsequent transmissions and transmits a feedback signal indicative of that determination, an implementation of the time diversity resulting in a better robustness to counter signal fading than if the spatial diversity were implemented and an implementation of spatial diversity resulting in an increase in a rate of data packet transfer over that if the time diversity were implemented, because the OFDM signals that are transmitted over multiple ones of the transmitters are independent of each other for the spatial diversity and correspond to each other for the time diversity, the controllers associated with the reception directing that transmission of at least one feedback signal occur that reflects the determination, the controllers including those associated with the transmission that respond to receipt of the feedback signal to direct an encoder to assign constellation points to the sub-carriers in accordance with a channel condition so as to classify each of the sub-carriers into one of two groups, the encoder including a space time transmitter diversity (STTD) encoder and a spatial multiplexing (SM) encoder, the STTD encoder being arranged to encode the sub-carriers classified in one of the groups in accordance with the time diversity and the SM encoder being arranged to encode the sub-carriers classified in the other of the groups in accordance with the spatial diversity.

30. A method as in claim 29, wherein the controllers associated with the reception make a calculation of eigen values of channel matrices to make a determination as to which sub-carriers are to use the time diversity to reduce signal fading forward error correction (FEC) during a subsequent transmission and which sub-carriers are to use the spatial diversity to increase a rate of data transfer during the subsequent transmission, the controllers associated with the reception make the determination based on a comparison between a threshold and at least one of three criteria and to direct transmission of a feed back signal indicative of a result of the determination, at least one of the criteria being based on the calculation, at least another of the criteria being based on elements of a diagonal of at least one of the channel matrices.

31. A method as in claim 29, wherein the controllers associated with the reception make the determination based on a comparison of a channel condition with a threshold, the channel condition being based on a frequency response channel matrix that is derived from OFDM symbols.

32. A method as in claim 31, further comprising calculating a smallest eigen value of the frequency response channel matrix basing the channel condition on the calculating.

33. A method as in claim 31, further comprising determining a smallest element in a diagonal of the frequency response channel matrix and basing the channel condition on the determining.

34. A method as in claim 31, further comprising calculating a ratio of largest and smallest eigen values of the channel matrix and basing the channel condition on the ratio.

35. A method as in claim 31, further comprising basing the channel condition on one of three criteria selected from a group consisting of a calculation of smallest eigen values of the channel matrix, a smallest element in a diagonal of the channel matrix, and a ratio of largest and smallest eigen values of the channel matrix.

36. A method as in claim 29, further comprising determining a modulation scheme on each of the sub-carriers based on an estimated ratio selected from a further group consisting of a carrier to interference ratio and a signal to noise ratio.

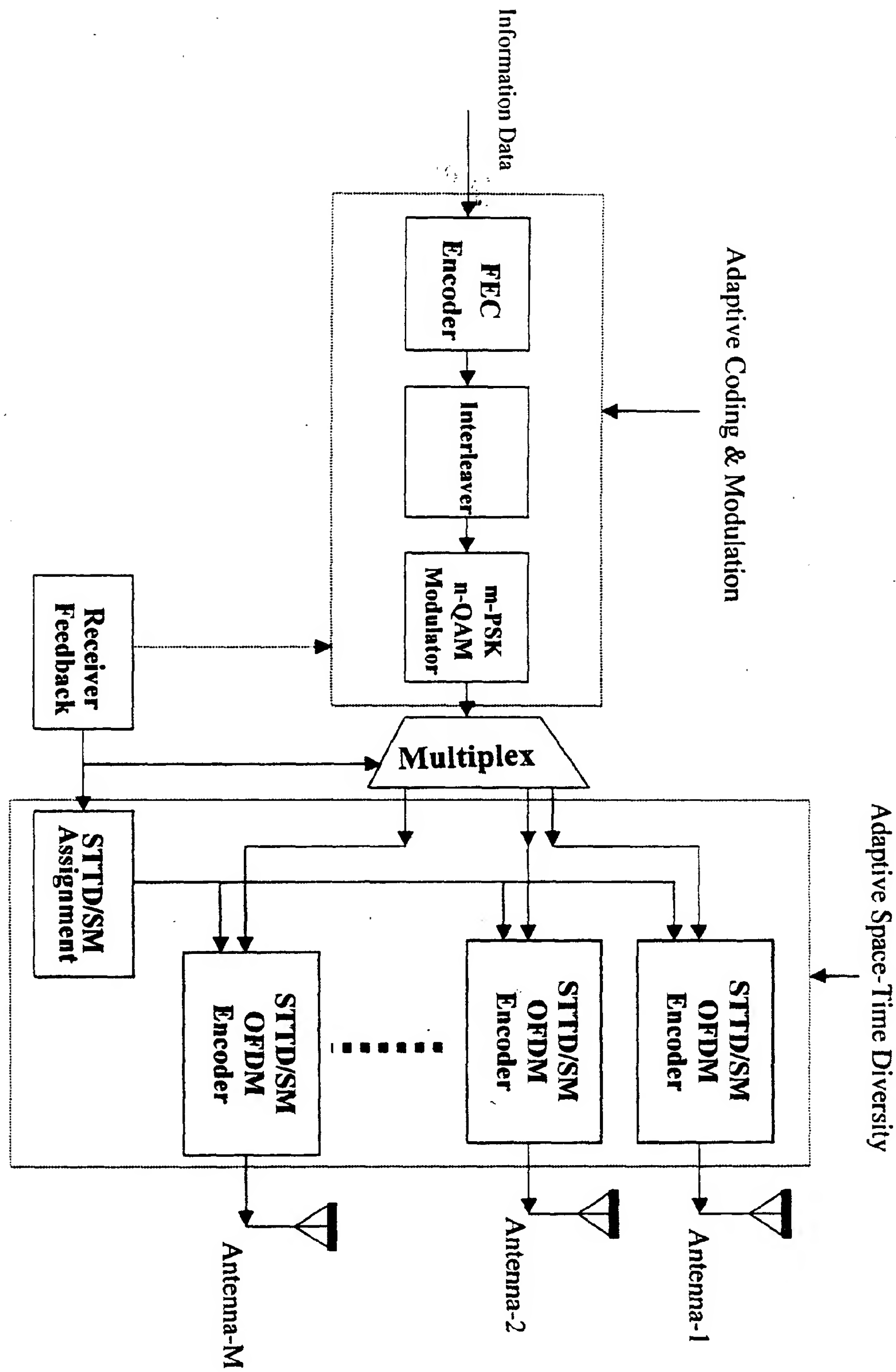

Information Data
FEC Encoder
Interleaver
m-PSK n-QAM Modulator
Adaptive Coding & Modulation
Receiver Feedback
Multiplex
STTD/SM Assignment
STTD/SM OFDM Encoder
STTD/SM OFDM Encoder
STTD/SM OFDM Encoder
Adaptive Space-Time Diversity
Antenna-1
Antenna-2
Antenna-M

Figure-1

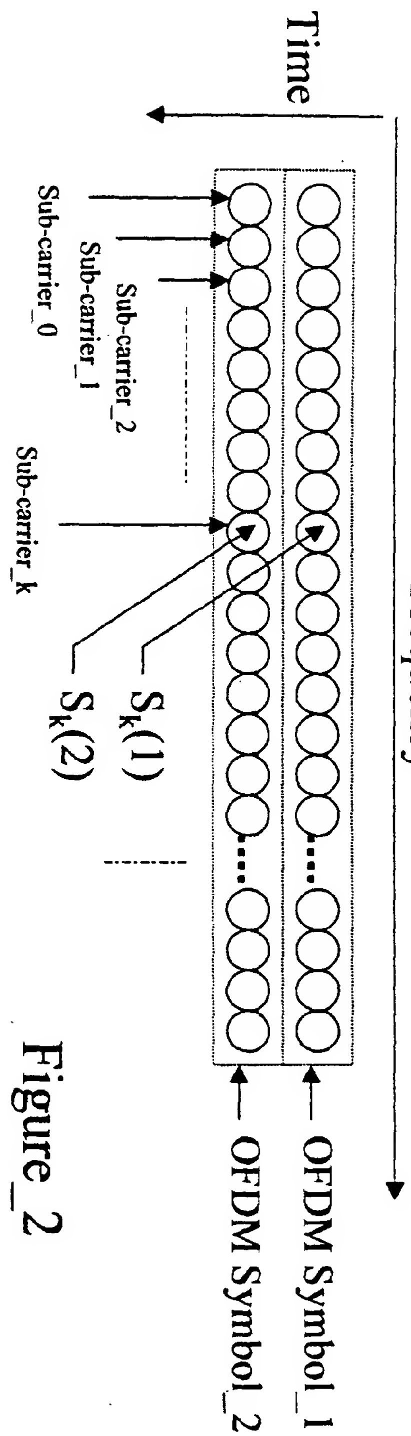


Figure_2

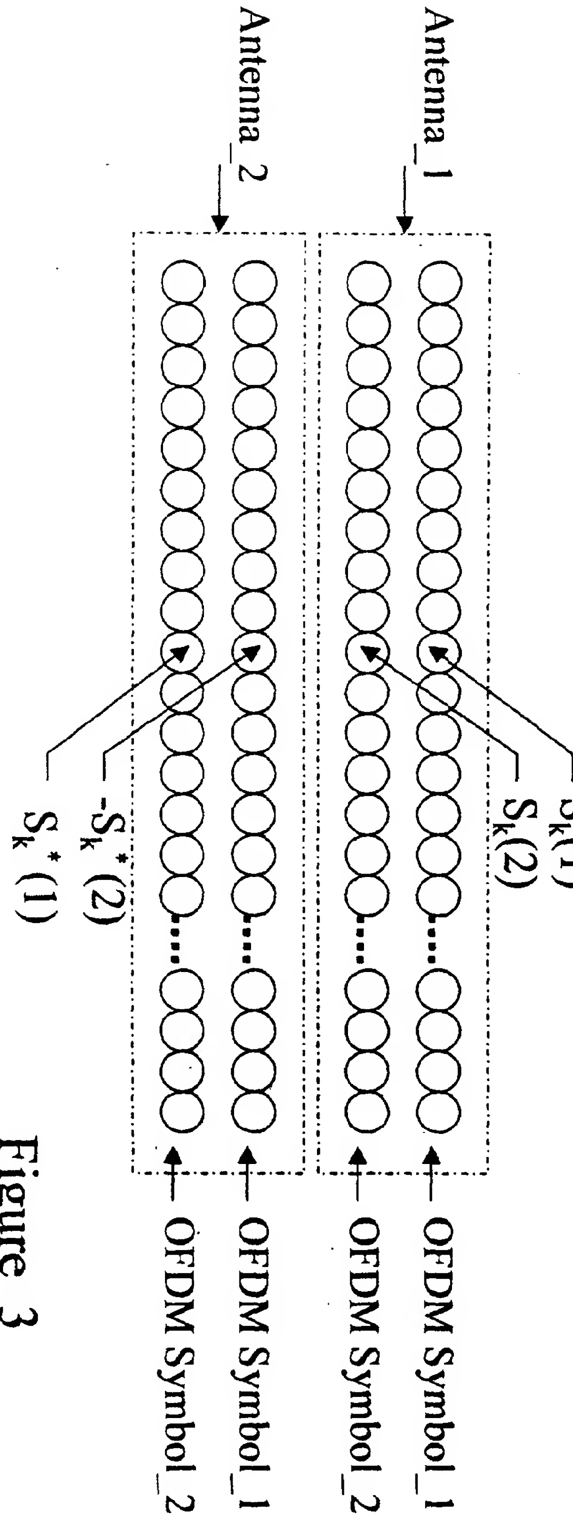


Figure_3

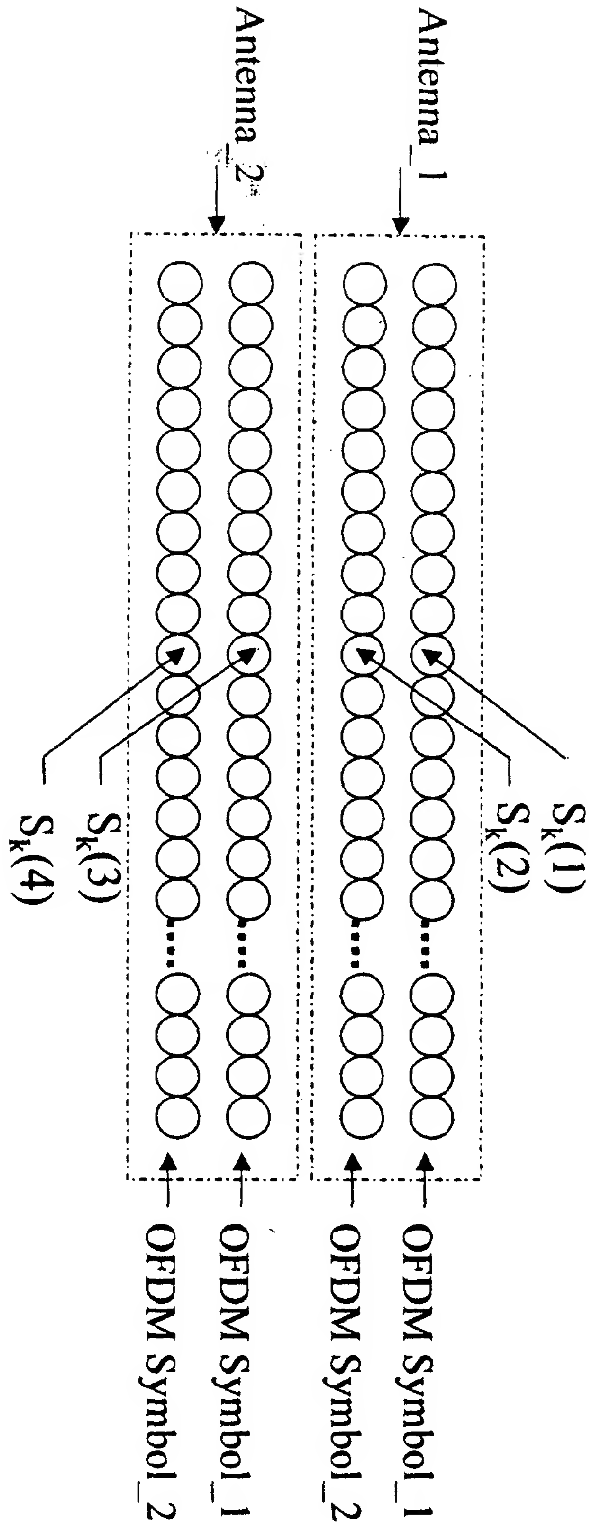
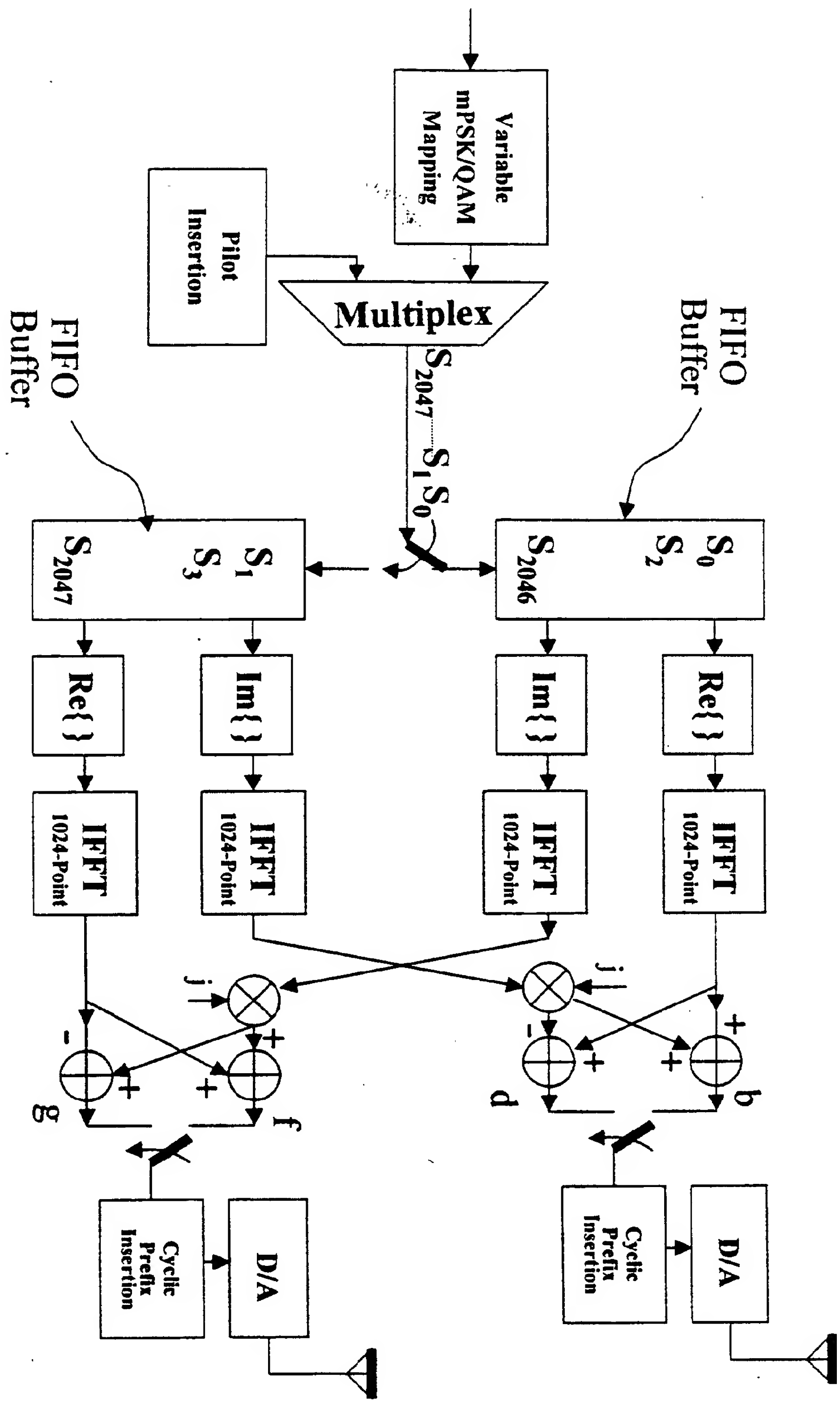

Antenna_1
Antenna_2
$S_k(1)$
$S_k(2)$
$S_k(3)$
$S_k(4)$
OFDM Symbol_1
OFDM Symbol_2
OFDM Symbol_1
OFDM Symbol_2

Figure_4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mPSK/QAM
Mapping
Pilot
Insertion
Multiplex
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FIFO
Buffer
FIFO
Buffer
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S_2
S_{2046}
S_1
S_3
S_{2047}
Re{ }
Im{ }
Im{ }
Re{ }
IFFT
1024-Point
IFFT
1024-Point
IFFT
1024-Point
IFFT
1024-Point
j
j
+
+
+
+
-
+
+
+
-
b
d
f
g
Cyclic
Prefix
Insertion
D/A
Cyclic
Prefix
Insertion
D/A

Figure_5

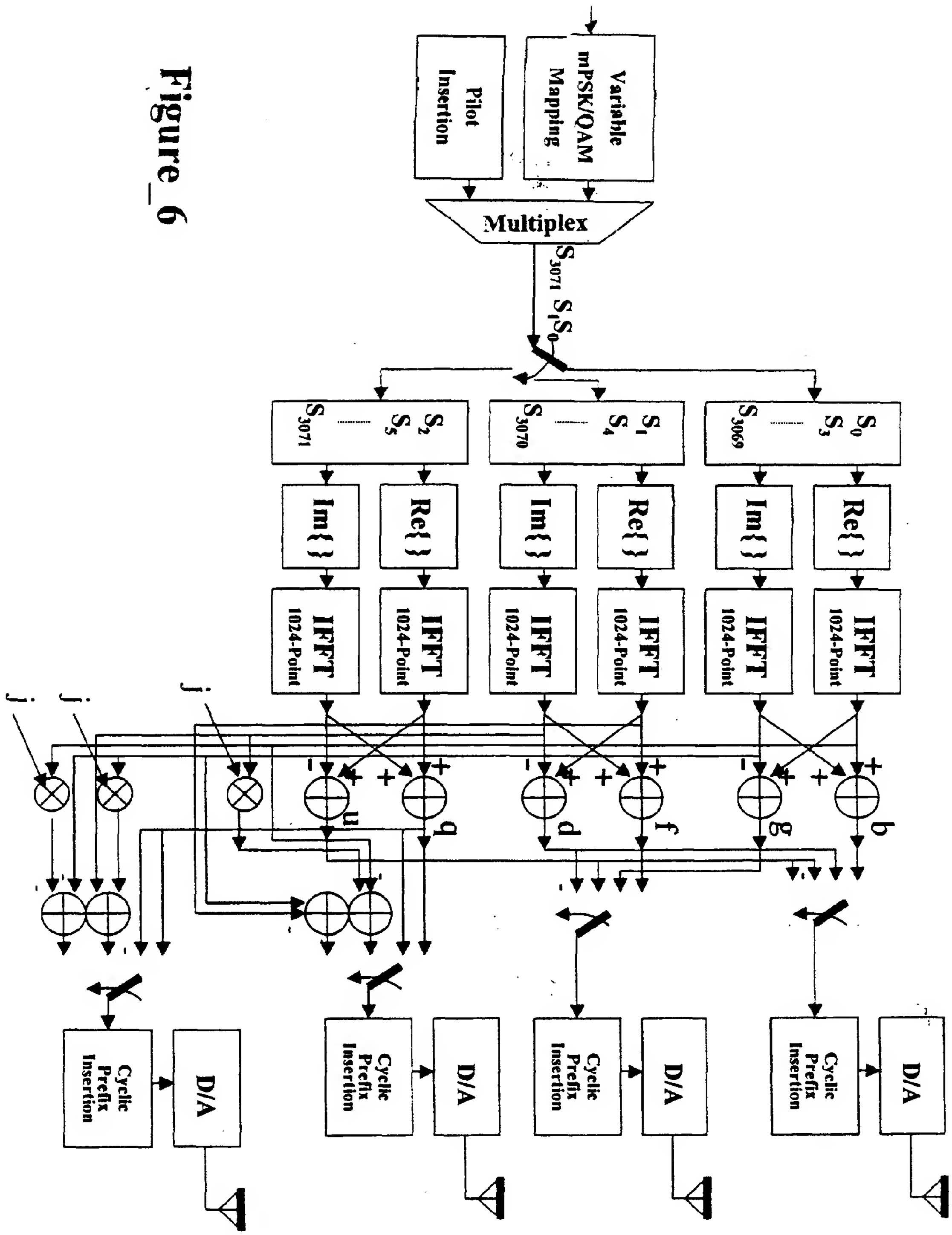
Variable
mPSK/QAM
Mapping
Pilot
Insertion
Multiplex
S_{3071} $S_1 S_0$
S_0 S_3 S_{3069}
S_1 S_4 S_{3070}
S_2 S_5 S_{3071}
Re{ }
Im{ }
IFFT
1024-Point
j
b
g
f
d
q
u
Cyclic
Prefix
Insertion
D/A

Figure_6

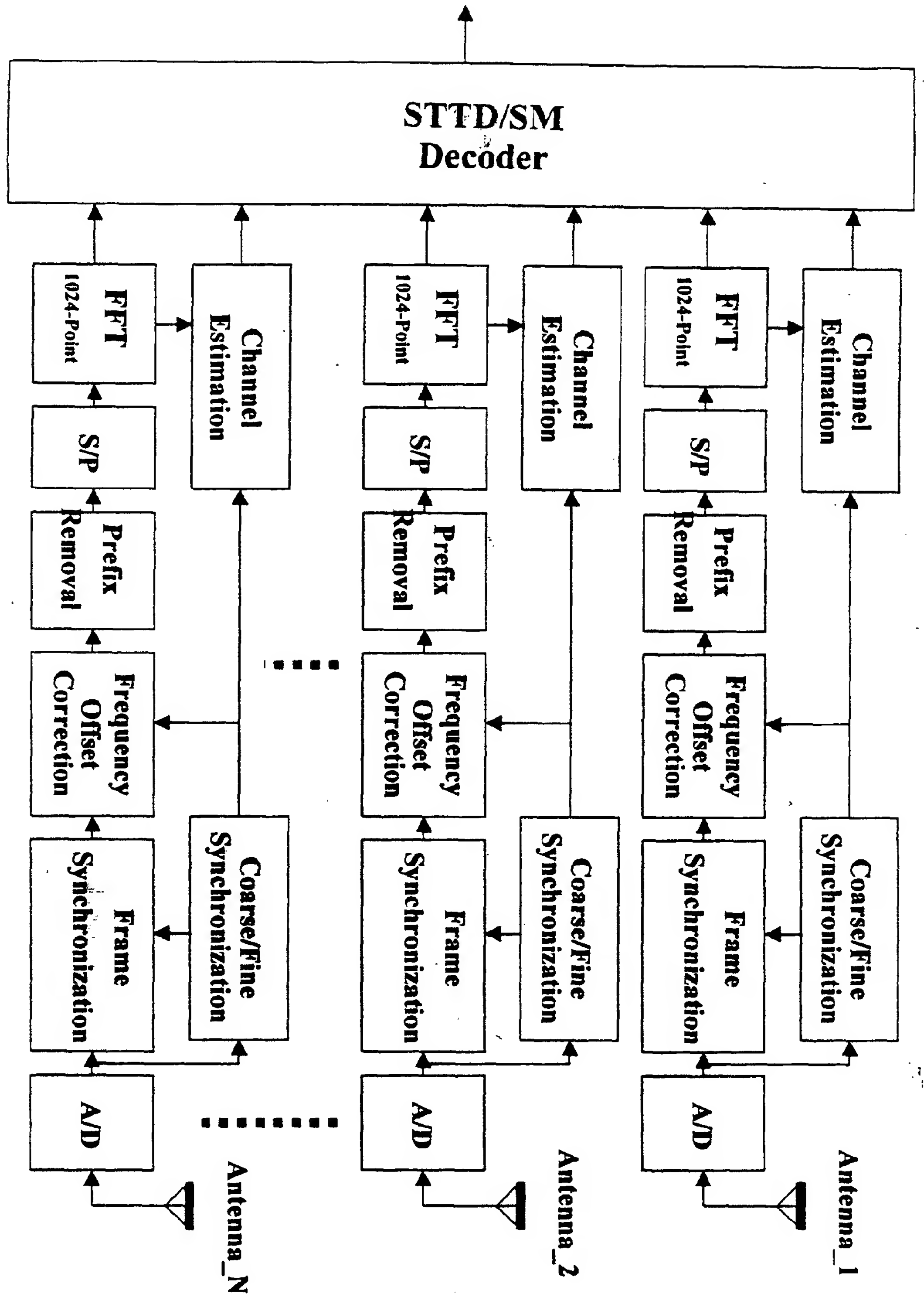
STTD/SM Decoder
FFT
1024-Point
Channel Estimation
S/P
Prefix Removal
Frequency Offset Correction
Frame Synchronization
Coarse/Fine Synchronization
A/D
Antenna_1
Antenna_2
Antenna_N

Figure_7

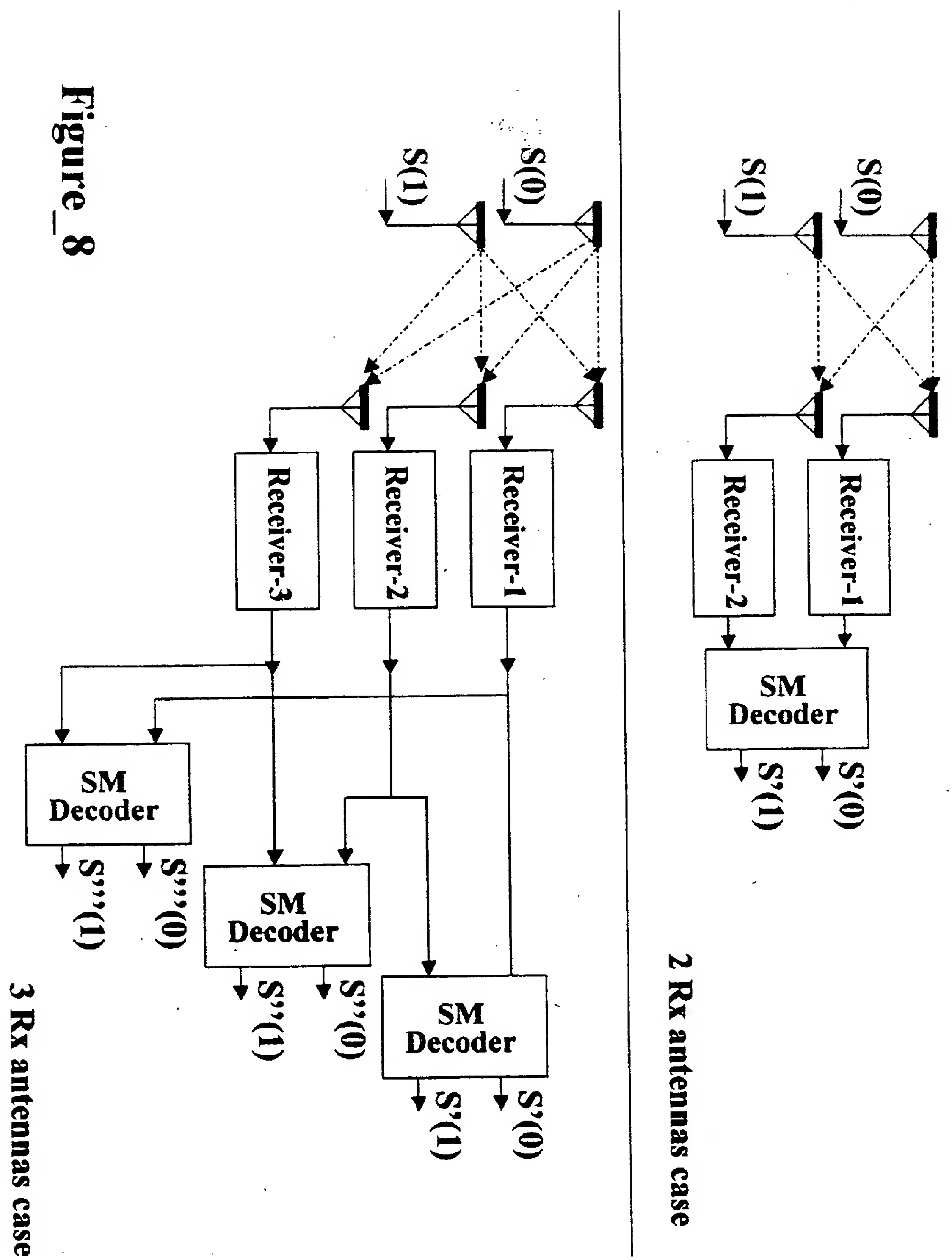
S(0)
S(1)
Receiver-1
Receiver-2
SM
Decoder
S'(0)
S'(1)
2 Rx antennas case
Receiver-3
S''(0)
S''(1)
S'''(0)
S'''(1)
3 Rx antennas case

Figure_8

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(54) **Method of managing time-slot interchange in transoceanic MS-spring networks**

(57) A method of managing time slot interchange in transoceanic MS-SP ring networks. The method, in case of ring failure in a single span of the path installed in a transoceanic MS-SP RING with Time Slot Interchange (TSI) mechanism, comprises the step of carrying out a ring switch action by the MS-SP mechanism, and is characterized by comprising the step of re-routing the path in the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span. The method according to the invention further provides for the managing of double-failure or multi-failure cases resulting in one or more nodes being isolated.

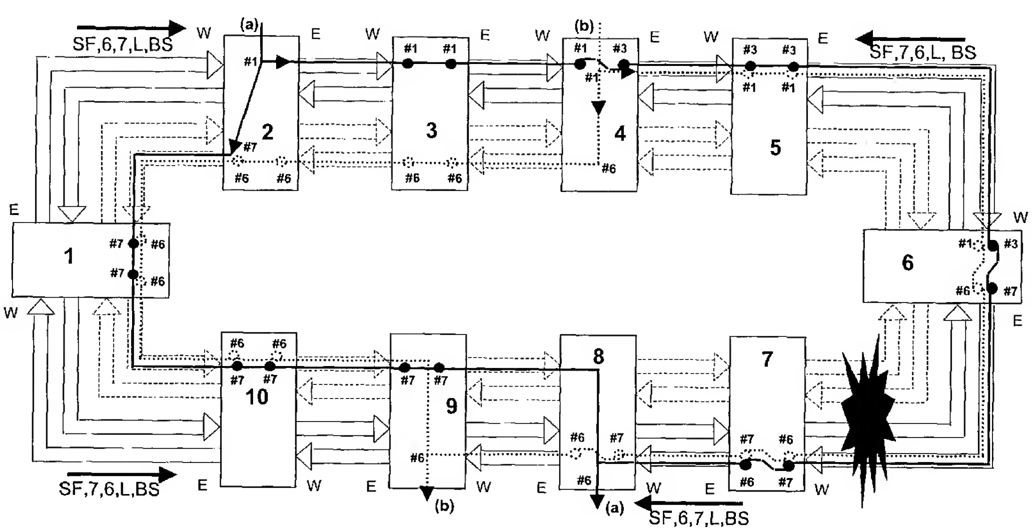

Fig. 3

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Description

[0001] The present invention relates to a method of managing changes of time-slot allocations in ring networks protected by a transoceanic MS-SPRING protection mechanism.

[0002] In SDH MS-SPRING (Multiplex Section Shared Protection Ring) networks, a shared protection mechanism is implemented, which mechanism allows for the automatic restoration of the traffic in case of defects or failures in the connection fibers. In other words, the MS shared protection ring networks perform the automatic restoration of the traffic through a synchronized re-routing of said traffic, performable at each node of the ring. This operation is controlled by a protocol consisting of messages that are continuously interchanged between adjacent nodes. Said protocol and the related operations are defined by several international standards issued by ANSI, ITU-T and ETSI and they are characterized by a set of rules and messages. See, for instance, Recommendation ITU-T G. 841.

[0003] Protection in an MS shared protection ring network is implemented according to a so-called Bridge and Switch technique that consists essentially in re-routing the traffic, by means of an appropriate modification in the internal connections of the network elements, switching it from the working capacity to the protection capacity. The protection in an MS shared protection ring network is a multiplex section-oriented protection mechanism, i.e. the events defining the traffic restoration and the hierarchy that regulates those events are given at multiplex section level. In the "classic" (or terrestrial) MS shared protection rings, in the event of a failure, the whole high-priority line capacity is re-routed by utilizing the corresponding low-priority line capacity; in the transoceanic MS shared protection rings, on the contrary, only the paths affected by a failure are selectively re-routed.

[0004] It is also known that the ring networks provide for a mechanisms termed "Time Slot Interchange", in short TSI. TSI means, for instance, that when traffic is configured in a given ring network, such a traffic, which is carried in the associated STM-n and hence in the AU-4 contained in the STM-n, is allowed to travel through a network element occupying different AU-4 numbers at the input and at the output. Consider for instance a maximum capacity of a four-fiber ring, composed of sixteen AU-4s. The TSI mechanism allows one to enter a network element (of pure transit and where no termination occurs) with AU-4#X from its West side (W) and to go out from its East side (E) with an AU-4#Y, where X≠Y = 1, 2, ..., 16. The advantage is a greater flexibility in the traffic allocation and a very efficient exploitation of the band.

[0005] At present, performing TSI in ring networks protected by an MS-SPRING protection mechanism is not known. In particular, it is not known to perform allocation changes in transoceanic MS shared protection ring networks.

[0006] Therefore the main object of the present invention is to provide a method allowing the execution of allocation changes in transoceanic rings protected by an MS-SPRING mechanism. This and further objects are achieved by a method having the features set forth in independent claim 1 and a network element according to claim 8. The respective dependent claims define further advantageous characteristics of the invention itself. All the claims are intended to be an integral part of the present description.

[0007] The basic idea of the present invention consists substantially in protecting the high-priority traffic by assigning, in case of a ring failure, the low-priority channel time slots chosen according to the real failure location and to the instant at which such failure has occurred, with respect to other failures possibly already present.

[0008] The invention will certainly become clear in view of the following detailed description, given by way of a mere non limiting and exemplifying example, wherein:

- Fig. 1 shows a ring network in a stable faultless situation, the network having a plurality of nodes, two installed paths and some allocation changes;
- Fig. 2 shows the same ring network of Fig. 1 just after a ring failure took place;
- Fig. 3 shows the ring network of Fig. 2 in a stable situation with a ring failure;
- Figs. 4 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of simultaneous double failure;
- Figs. 5 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of nearly simultaneous double failure;
- Figs. 6 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of double failure at different times (first sub-scenario);
- Figs. 7 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of double failure at different times (second sub-scenario);
- Figs. 8 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of clearing of a first failure; and
- Figs. 9 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of clearing of a second failure.

[0009] In the various figures, a four-fiber transoceanic telecommunication ring has been always depicted. The two working fibers (otherwise known as "high-priority channels" or "HP channels") are indicated by solid-line

arrows whereas the protection fibers (otherwise known as "low-priority channels" or "LP channels") are indicated by dashed-line arrows. Naturally, the present invention applies both to the illustrated case of bi-directional traffic and to the case of unidirectional traffic.

[0010] Moreover, the present invention is applicable also to rings in which the traffic subjected to TSI is configured with "channel concatenation (AU4)".

[0011] The ring illustrated to describe the invention comprises ten network elements or nodes, represented by blocks and designated by respective numerals (1 to 10). The West (W) and East (E) sides of each node are indicated. The term "span" is used throughout this description to mean that part between two adjacent nodes, for instance between nodes 1 and 2 or the one between nodes 7 and 8.

[0012] In the ring there are depicted, by way of a non-limiting example, two paths installed, "path (a)" and "path (b)". The first path (path a) is depicted by a bolt solid line whereas the second path (path b) is depicted by a bolt dotted line. Path (a) is inserted at node 2 and is dropped at node 8. Path (b) is inserted at node 4 and is dropped at node 9.

[0013] Finally, it has been tried to clearly indicate (with numbers after symbol "#") the various time slots in which the various paths are allocated, span by span. Thus it has been also indicated if a Time Slot Interchange (TSI) occurs at a node or if that node allows that path to transit without changing the AU-4 on which it is allocated.

[0014] The present invention contemplates the general criteria set forth below:

I) single failure: once a ring failure has occurred in a given span, a ring switch action is performed by the MS shared protection mechanism. This activity defines the set of re-routable paths, namely all the paths whose nominal route includes the failured span. According to the present invention, each of these paths is re-routed on the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span. There is no risk of any conflict since the LP time-slot assignment criterion is the same for all the failured paths.

II) double failure: if a failure occurs at a further span and the path can still be saved, then

II.I) i) the actual re-routing is released; ii) among the two failured spans one is chosen according to a certain criterion; and iii) the path is re-routed over the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the selected failured span. Should multiple (more than two) failures occur, the choice of the span to consider for the TSI path protection is to be made by selecting, according to the above criterion, among the spans adjacent to the switching nodes that are able to communicate with the termination nodes of the path to be protected. There is no risk of any conflict since the LP time-slot assignment criterion is the same for all the paths affected by the failure. In this way, any transient misconnection is avoided.

II.II) The actual re-routing is not released when the persistency of the re-routing information is supported by the ring network elements.

[0015] The procedures that are implemented by each node of the ring (in addition to the procedures already provided for by the MS shared protection mechanism) will be indicated below:

A. If, at both W and E sides of the node, two Bridge Requests with an "Idle" status code concerning the same span (single failure) are detected, then each path comprising the span in question is re-routed over the LP time slot equal to the HP time slot of the (sole) span affected by the failure. In other words, if the node is a path termination node (a node where the path to be re-routed is inserted or dropped), the Bridge & Switch action is performed by utilizing the LP time slot equal to the HP time slot of the span affected by the failure.

A1. If at W or E side of the node at least one Bridge Request with an "Idle" status code is detected, a pass-through is performed, if necessary, by utilizing the same LP time slot (LP time slot equal to the HP time slot of the span affected by the failure).

B. If at both W and E sides of the node Ridge Requests with a "Bridged and Switched" status code concerning different spans (multiple failures) are detected, then each path comprising the spans in question is re-routed over the LP time slot equal to the HP time slot of the span affected by the failure that has been chosen according to a fixed criterion. The criterion for selecting one among the failed spans could, for instance, be any one of the following:

a) the failed span adjacent to the switching node with higher (or lower) node identification is chosen;

b) the failed span adjacent to the switching node coming first (or last) in the ring map; or

c) the failed span adjacent to the switching node that is "far West" or "far East" node in the ring.

[0016] Similarly to the above case A, if the node in question is a path termination node (node in which the path to be re-routed is inserted or dropped), the Bridge & Switch action is performed by utilizing the LP time slot equal to the HP time slot of the selected failured span.

B1. If at the W or E side of the node at least one

Bridge Request with "Bridged and Switched" status code is detected, a pass-through is performed, if necessary, by utilizing the same LP time slot as above.

C. In an optimized embodiment, should a path re-routing due to single failure be under way, if at W and E sides of the node Bridge Requests with a different ("Idle" or "Bridged and Switched") status code that carry the indication of a second failure, hence located at a different span, are detected, then, for each path that has already been re-routed, it is evaluated if the new requests exhibit a failured state that requires to release or maintain the present re-routing. It is not necessary to release the actual re-routing of a single path in the following cases:

i) when a failure is detected in addition to the already existing failure/s (and such new failure/s does/do not result in the "isolation" condition of any of the termination nodes of the already protected path); or ii) when the last occurred span failure has been removed.

[0017] It is understood how the persistency of the path re-routing condition is strictly connected to the coexistence of the aforesaid triggers at both sides of the nodes. This behavior results in the correct re-routing of the failured paths, namely it prevents misconnections from being created during transient states of the ring protection mechanism.

[0018] Note that the above is applicable to the case of bi-directional path and unidirectional one not using the inverse route. Clearly, if two unidirectional paths are allocated on the same time slot (in opposite directions), then the same LP time slot can be assigned to both paths.

[0019] Fig. 1 shows a transoceanic MS shared protection ring protected against failures in accordance with the invention, at some nodes of the ring Time-Slot Interchange (TSI) occurring. The installed paths are two: (a) and (b). Path (a) enters the ring at node 2 and is allocated on the AU-4 #1; therefore, in span 2-3 the allocation is AU-4#1; at node 3 the allocation is unchanged (therefore it remains AU-4#1); at node 4, the allocation is changed from AU-4#1 to AU-4#3; in span 4-5 the allocation is therefore AU-4#3; at node 5 the allocation is not changed (therefore it remains AU-4#3); in span 5-6 the allocation is therefore AU-4#3; at node 6 the allocation is changed from AU-4#3 to AU-4#7; therefore, in span 6-7 the allocation is AU-4#7; at node 7 the allocation changes again from AU-4#7 to AU-4#6; therefore, in span 7-8 the allocation is AU-4#6; finally, path (a) is dropped at node 8. For path (b): it enters at node 4 and is allocated on AU-4#1; this allocation is maintained up to node 6 where it changes from AU-4#1 to AU-4#6; it is changed again at node 7 (passing from AU-4#6 to AU-4#7) and at node 8 (passing from AU-4#7 to AU-4#6). Finally, path (b) is dropped at node 9.

[0020] In the event of a ring failure (namely, a failure that makes both high-priority (HP) channels and low-priority (LP) channels useless), the present invention provides for a method of remedying such failure notwithstanding the presence of allocation changes in the ring. Reference should be made, for the event of single ring failure, to Figs 2 and 3 where a ring failure in the span 6-7 has been simulated.

[0021] As it is known, the management of failures in the synchronous (SDH or SONET) telecommunications networks occurs, for some protection types among which the MS-SPRING one, through bytes K1 and K2 of the frame overhead. Since the present invention does not concern specifically such bytes K1 and K2, a more precise description thereof will not be given, the reader having the possibility to refer to relevant Recommendations, for instance the ITU-T Recommendation G. 841, incorporated herein by reference.

[0022] In the event of a failure, the nodes (6 and 7) that are adjacent to the failure will send, as usual, proper failure signaling in the direction opposite to the failure. The structure of the request (APS signaling) is the following: Bridge Request, Destination Node ID, Source Node ID, Type of Path, Protection Status. In this instance, the node 6 will send a signaling of the type SF, 7,6,L,ID (Signal Fail, Destination Node: 7, Source Node: 6, Path: Long, Protection Status: Idle) to indicate that a ring failure occurred at span 6-7 and that no action has been taken for the time being. Node 7 will do the same by sending a signaling type SF,6,7,L,ID from its East side (E).

[0023] Such signaling will travel down the ring in opposite directions and will be received by termination nodes (2, 8; 4, 9) of paths (a) and (b) that will perform the requested Bridge and Switch (BR & SW) by utilizing the LP channels. In accordance with the present invention, the ring protection (BR&SW and pass-through), in the event of single ring failure, is performed by squelching the terminations of LP traffic allocated on the AU-4s corresponding to the failured span and by allocating the HP traffic on such AU-4s. With reference to Fig. 3, since in span 6-7 the path (a) was allocated on the HP AU-4#7 and the path (b) was allocated on the HP AU-4#6, the allocation AU-4#7 (of the LP channels) will be utilized for the first path and the allocation AU-4#6 (of the LP channels) will be utilized for the second path.

[0024] Should TSI be configured also on low-priority traffic, the high-priority traffic protection that requires the use of one of the LP channels utilized in the low-priority TSI, will anyway result in the squelching action on both the low-priority traffic terminations.

[0025] Once a node adjacent to the failure has received the signaling sent by its homologous opposite side, with Protection Status corresponding to "Idle", the node itself will send a modified signaling (with Protection Status = Bridged & Switched, BS). In other words, node 6 will send SF,7,6,L,BS from its West side whereas node

7 will send SF,6,7,L,BS from its East side. Upon restoration of the full ring functionality (fault clearing) the BR&SW will be removed and the failure signaling (SF, 7,6,L,BS and SF,6,7,L,BS) will be removed.

[0026] The present invention, in addition to the single failure event illustrated above, provides for a traffic protection method applicable to multiple failures leading to isolation of one or more nodes in which the TSI of the installed path/s is configured. Within this context three failure scenarios are considered and separately described: in the first scenario the failures occur simultaneously, in the second scenario the failures occur nearly simultaneously while in the third scenario the failures occur at different times.

[0027] Referring initially to Figs. 4, consider the case where two failures (SF1 and SF2) occur exactly at the same time instant. For simplicity, paths (a) and (b) before the occurrence of the failures, are allocated in a manner similar to what described for Fig. 1 and therefore the description of the allocations will not be repeated here. Upon the occurrence of the first failure (SF1) on the span 6-7, the node 6 (Fig. 4.1) will send a failure signaling (SF,7,6,L,ID) from the West side whereas, upon the second failure (SF2) on the span 7-8, node 8 will send a simultaneous failure signaling (SF,7,8,L,ID) from the East side (Fig. 4.2).

[0028] At the time when each of the two signaling with "Idle" code, which were generated by the switching nodes, is received by the termination nodes of the paths to be protected, squelching of the local termination (if any) of the LP channel corresponding to the HP channel allocated in the failed span to which the signaling is referred, takes place; while, at the nodes designed to realize the pass-through of the LP channels, the squelching of the local termination (if any) of the LP channel corresponding to the HP channel allocated in the failed span to which the signaling is referred takes place and the subsequent pass-through connection also takes place. The actions just described (squelching and squelching + pass-through) are removed both from the path termination nodes and from the pass-through nodes, as soon as such nodes receive the second signaling generated by the switching nodes.

[0029] When (Fig. 4.3) the signaling (SF,7,6,L,ID) containing the "Idle" code of SF1 reaches node 8, node 8 (Fig. 4.4) will send a signaling containing BR&SW (BS) Status Code of the type SF,7,8,L,BS. The same will be for node 6 (Fig. 4.5): as soon as it receives signaling (SF,7,8,L,ID) containing the "Idle" code of SF2, it will send a signaling containing BR&SW Status Code (BS) of the type SF,7,6,L,BS.

[0030] At the time when one of the two signalings with BS code generated by switching nodes is received by the termination nodes of the HP paths to be protected, the squelching of the local termination (if any) of the LP channel to be used for the protection, that was chosen according to one of the criteria described above, takes place; while, at the nodes designed to realize the pass-through of the LP channels, the squelching of the local termination (if any) of the same LP channel will take place and also the subsequent pass-through connection, will take place.

[0031] The Bridge & Switch action that is performed on the LP channel chosen according to the same criterion as above, is performed by every termination node of the HP paths to be protected, as soon as both signalings with BS Code (SF,7,6,L,BS and SF,7,8,L,BS) are detected at the two sides of the node itself.

[0032] Thus, a stable state of the protected ring has been achieved.

[0033] Referring initially to Figs. 5, consider the case where two failures (SF1 and SF2) occur nearly at the same time instant (or anyway failure SF2 occurs before the situation following SF1 is stabilized). For simplicity, paths (a) and (b) before the occurrence of the failures, are allocated similarly to what described for Fig. 1 and therefore the description of the allocations will not be repeated here. Upon the occurrence of the first failure (SF1) in span 6-7, the node 6 will send a failure signaling (SF,7,6,L,ID) from the West side and, similarly, it will send another failure signaling (SF,6,7,L,ID) from East side. See Figs. 5.1 and 5.2.

[0034] Suppose (Fig. 5.3) that the failure signaling (SF,6,7,L,ID) from the East side is able to reach node 8 before the second failure (SF2) occurs in span 7-8, which results in node 7 isolated. Upon the second failure (SF2), node 8 (node adjacent to the failure) will send a corresponding failure signaling (SF,7,8,L,ID) from its East side. Anyway, the signaling of the second failure will follow the first failure one (Fig. 5.4).

[0035] As soon as the signaling (SF,6,7,L,ID; SF, 7,6,L,ID) containing the "Idle" code of the first failure reach the termination node 2 (Fig. 5.5) of path (a), this node will perform the BR&SW action by utilizing the LP AU-4 corresponding to the span affected by the first failure (LP AU-4#7, in this instance). However, as soon as also the new signaling (SF,7,8,L,ID) of the second failure (SF2) reaches node 2, the BR&SW action, just realized, is removed (Fig. 5.6).

[0036] Analogously (Fig. 5.7), as soon as the signaling (SF,6,7,L,ID; SF,7,6,L,ID) containing the "Idle" code of the first failure reach the termination node 4 of the path (b), this node will perform the BR&SW action by utilizing the LP AU-4 corresponding to the span affected by the first failure (LP AU-4#6 in this case). However, as soon as also the new signaling (SF,7,8,L,ID) relating to the second failure (SF2) reaches node 4, the BR&SW action, just realized, is removed (Fig. 5.8).

[0037] Obviously, the actions to be taken before the just described temporary BR&SWs are the squelching of the local termination (if any) of the LP channel associated with the span 6-7 both on the termination nodes of the paths to be protected, and on the nodes designed to realize the pass-through, as well as the pass-through connection of the LP channel itself: in order a node to perform such actions, the reception of at least one of the

two signaling with "Idle" code generated by the switching nodes is enough to the interested node.

[0038] At the same time, when the signaling (SF,7,6,L, ID) containing the "Idle" code of the first failure reaches node 8, node 8 will send a signaling containing the BR&SW (BS) status code of the type SF,7,8,L,BS (Fig. 5.9). The same will be for node 6: as soon as it receives the signaling (SF,6,7,L,ID) containing the "Idle" code of the first failure, it will send a signaling containing the BR&SW (BS) Status Code of the type SF,7,6,L,BS (Fig. 5.10).

[0039] Because of the presence of the new signaling (SF,7,8,L,ID) concerning the second failure (SF2), node 6 will change again its signaling from SF,7,6,L,BS to SF, 7,6,L,ID (Fig. 5.11).

[0040] At this stage both the signaling transmitted by node 8 containing the BR&SW Status Code and the two consecutive signaling transmitted by node 6 respectively containing the BS&SW (BS) and the "Idle" status codes are present in the ring. The signaling containing the BS Status Code result, at the nodes detecting them, in the squelching of the local termination (if any) of the LP channel chosen for the protection according to one of the aforesaid criteria (for instance the LP channel corresponding to the allocation used in the span affected by the first failure, AU-4#6), as well as in the pass-through of such LP channel at the nodes designed to perform such a function. It is to be noted that, among the two signaling that consecutively emitted by node 6, the one containing "Idle" code does not remove the squelching and pass-through actions activated by the previous signaling (with BS code), since both refer to the same failed span: SF,7,6,L,BS and SF,7,6,L,ID.

[0041] The node 9 (Fig. 5.12), receiving a signaling with BS code (SF,7,6,L,BS and SF,7,8,L,BS) from both its W and E sides, will perform the BR&SW action by utilizing the LP allocation related to one of the failured spans, for instance the one affected by the first failure (AU-4#6). Node 8, that receives the signaling containing the BS code (SF,7,6,L,BS) previously sent to it by node 6, will realize the BR&SW action (Fig. 5.13) by utilizing the LP allocation related to one of the failured spans, for instance the one affected by the first failure (AU-4#7). Some of the possible selection criteria have been mentioned above.

[0042] Since the request that reaches both node 9 and node 8 with "Idle" code, is related to the failured span already indicated in the preceding request (SF, 7,6,L,BS), the BR&SW action is maintained (Fig. 5.14).

[0043] When the request related to the second failure (SF2) and containing the BS code reaches node 6, the APS signaling is updated with the BS code, namely node 6 will send, from side W, the signaling SF,7,6,L,BS (Fig. 5.15).

[0044] Node 4, as soon as it receives signaling with BS code from both sides, will realize the BR&SW action by utilizing the LP allocation related to one of the spans affected by a failure, for instance the one affected by the first failure (AU-4#6).

[0045] Lastly, the node 2, as soon as it receives, from both its W and E sides, a signaling with BS code (SF, 7,6,L,BS and SF,7,8,L,BS), will perform the BR&SW action (Fig. 5.16) by utilizing the LP allocation related to one of the spans affected by a failure, for instance the one affected by the first failure (AU-4#7).

[0046] Thus, a stable state in the protected ring is obtained.

[0047] As said above, the scenario of the actions taken by the various nodes is different in the case where the failures do not occur at the same time. In this connection, two different sub-scenarios should be distinguished. With reference to Figs 1 to 3 and 6, the actions and the consequences related to the first sub-scenario are schematically listed below starting from a situation free of faults.

[0048] The first failure (SF1) occurs. The node 6 sends SF,7,6,L,ID from the W side. The node 7 sends SF,6,7,L,ID from the side E (Fig. 2).

[0049] SF,7,6,L,ID and SF,6,7,L,ID reach the termination nodes of paths (a) and (b). The termination nodes perform the BR&SW action for each path to be protected by utilizing the corresponding LP channels of the span affected by SF1. Path (a) is allocated on LP AU-4#7. Path (b) is allocated on LP AU-4#6 (Fig. 3).

[0050] The nodes 6 and 7 adjacent to the failure SF1 send respective signaling with BS code (SF,7,6,L,BS and SF,6,7,L,BS) and a stable scenario of ring protected against SF1 is obtained (Fig. 3).

[0051] SF2 occurs on span 7-8: node 7 is isolated (Fig. 6.1). Node 8 sends SF,7,8,L,ID from the side E (Fig. 6.2).

[0052] The BR&SW action (both the BR&SW and the pass-through at the intermediate nodes) performed for path (a) is removed (Figs. 6.3 and 6.4). The BR&SW action (both the BS&SW and the pass-through at the intermediate nodes) performed for path (b) is removed (Figs. 6.5 and 6.6).

[0053] Node 6 receives the signaling SF,7,8,L,ID and sends SF,7,6,L,ID (Fig. 6.7).

[0054] The node 8 receives from node 6 the signaling SF,7,6,L,ID and sends SF,7,8,L,BS (Fig. 6.8).

[0055] Node 6 receives the signaling SF,7,8,L,BS and sends SF,7,6,L,BS (Fig. 6.9).

[0056] Nodes 2 and 8 receive the signaling SF,7,8,L, BS and SF7,6,L,BS and perform the BR&SW action by utilizing for instance the LP channels with AU-4 corresponding to that of the first failed span (LP AU-4#7). The scenario becomes stable for the path (a) (Figs. 6.10 and 6.11).

[0057] Nodes 4 and 9 receive the signaling SF,7,8,L, BS and SF,7,6,L,BS and perform the BR&SW action by utilizing for instance the LP channels with AU-4 corresponding to the one of the first failed span (LP AU-4#6). The scenario becomes stable for path (b) (Figs. 6.12, 6.13).

[0058] The squelching actions of the local termination

(if any) of the LP channel chosen for the protection according to one of the criteria already described and the subsequent pass-through of the same LP channel at the intermediate nodes come before the BR&SW actions just described and are performed with the rules already pointed out for the two previous scenarios.

[0059] With reference to Figs. 1 to 3 and 7, the actions and the consequences related to the second sub-scenario of double failure at different times will now be schematically listed in the following, still starting from a faultless situation.

[0060] The first failure (SF1) occurs. Node 6 sends SF,7,6,L,ID from the W side. Node 7 sends SF,6,7,L,ID from the E side (Fig. 1).

[0061] SF,7,6,L,ID and SF,6,7,L,ID reach the termination nodes of the paths (a) and (b). The termination nodes perform the BR&SW action for each path to be protected by utilizing the corresponding LP channels of the span affected by SF1. The path (a) is allocated on LP AU-4#7. The path (b) is allocated on LP AU-4#6 (Fig. 2).

[0062] The nodes 6 and 7 adjacent to failure SF1 receive the signaling with ID code (SF,7,6,L,ID and SF, 6,7,L,ID), send respective signaling with BS code (SF, 7,6,L,BS and SF,6,7,L,BS) and a stable scenario of ring protected against SF1 is obtained (Fig. 3).

[0063] SF2 occurs in span 7-8: node 7 is isolated (Fig. 7.1). Node 8 sends SF,7,8,L,ID from the E side (Fig. 7.2).

[0064] The node 8, as a node adjacent to the failure and as a termination node, evaluates whether the already protected paths can still be protected. In the affirmative, no action is taken; in the negative, the BR&SW action is removed (Fig. 7.3).

[0065] Node 9 receives the SF,7,8,L,ID request and evaluates whether the already protected paths can still be protected. In the affirmative, no action is taken; in the negative, the BR&SW action is removed (Fig. 7.4).

[0066] Node 2 receives the SF,7,8,L,ID request and evaluates whether the already protected paths can still be protected. In the affirmative, no action is taken; in the negative, the BR&SW action is removed (Fig. 7.5).

[0067] Node 4 receives the SF,7,8,L,ID request and evaluates whether the already protected paths can still be protected. If yes, no action is taken; if not, the BR&SW action is removed (Fig. 7.6).

[0068] As soon as node 6 receives the SF,7,8,L,ID signaling, it updates its request by inserting the "Idle" status code, of the type SF,7,6,L,ID. After a further signaling exchange, the nodes adjacent to the failure update the respective signaling by inserting "BR&SW" (BS) status code.

[0069] At this point only signaling with BS code are traveling in the ring that is failured by SF1 and SF2 and therefore a stable scenario for paths (a) and (b) towards the failures SF1 and SF2 has been achieved.

[0070] It will be recognized that the first sub-scenario results in a rather simple implementation since it is not necessary to store the failure "history" but, at the same time, traffic is not safeguarded in an optimal manner because the BR&SW is always removed. On the contrary, the second sub-scenario safeguards the traffic in a better manner but it is more difficult to be implemented because the traffic "history" shall be stored.

[0071] Having analyzed in detail the single-failure and the double-failure situations (simultaneous, nearly simultaneous or at different times), we will go on in describing schematically the actions that each network node must perform (and the corresponding consequences) when the failures are cleared and the ring functionality is restored.

[0072] Start from a stable situation of two failures SF1, SF2: in this situation, node 7 is isolated (Fig. 8.1) and only signalings with BS code (SF,7,8,L,BS and SF,7,6,L, BS) are traveling in the ring. Consider to clear first SF1: the node 7, no longer isolated, begins to send the APS signaling with "Idle" code related to the span affected by a failure (SF2) still present between the nodes 8 and 7 (SF,8,7,L,ID) (Fig. 8.2).

[0073] Since the LP allocation of the span 7-6 had been chosen, the BR&SW (and squelching of the any local termination of the LP channel utilized) action at node 4 must be removed. Similarly, as soon as also the SF,8,7,L,ID signaling reaches the other path termination nodes (2, 9, 8), the BR&SW and any local squelching action is removed also at such nodes 2, 9, 8 (Figs. 8.3 to 8.5). The removal of "BR&SW" at the termination nodes is accompanied by the removal of the pass-through (and of any local squelching) from the intermediate nodes that have performed the pass-through of the LP channel heretofore utilized for the protection. Since the signalings present at the intermediate nodes are related to the same span affected by a failure, such nodes can perform, if required, the pass-through of the LP channels, related to the current failure, to be utilized for the path protection.

[0074] The node 8, as a node adjacent to the failure SF2, receives SF,8,7,L,ID and changes the code of its signaling from SF,7,8,L,BS to SF,7,8,L,ID (Fig. 8.6). Such signaling with ID code gradually reaches all the termination nodes (9, 2, 4) showing them in this way that a single failure (SF2) is present. The termination nodes in turn will execute the BR&SW action (Figs. 8.8 to 8.10) by utilizing the LP channels that correspond to the failured span (for path (a) the LP AU-4#6 will be utilized, for path (b) the LP AU-4#7 will be utilized).

[0075] The nodes (7, 8) adjacent to the failure still present (SF2) will send corresponding signaling with BS code (SF,8,7,L,BS and SF,7,8,L,BS) and a single-failure stable condition will be achieved (Figs. 8.11, 8.12).

[0076] As soon as also SF2 is cleared, the ring will reach the faultless stable condition (Figs. 8.13, 8.14), with the progressive removal of the "Bridge" and "Switch" actions from all the path termination nodes and the consequent signalings with "No Request, Idle" code (NR,9,8,S,ID and NR,6,7,S,ID) by all the ring nodes, including nodes (7, 8) adjacent to the just cleared failure

(SF2).

[0077] Start now from a stable situation of two failures SF1, SF2 (Fig. 9.1): in this situation node 7 is isolated and only signalings with BS code (SF,7,8,L,BS and SF, 7,6,L,BS) are traveling in the ring. Consider to clear SF2 first: node 7 (Fig 9.2), no longer isolated, begins to send the APS signaling with "Idle" code (SF,6,7,L,ID) related to the span affected by failure (SF1) still present between nodes 6 and 7.

[0078] Since just the LP allocation of the span 7-6 had been chosen, the BR&SW action at node 8 can be maintained (Fig. 9.3). Similarly, the SF,6,7,L,ID signaling reaches the other path termination nodes (9, 2, 4) but the BR&SW action is maintained also at such nodes 9, 2, 4 (Figs. 9.4 to 9.6).

[0079] The same processing is carried out at intermediate nodes that perform the pass-through of the LP channels used for the protection: the pass-through is maintained.

[0080] Finally, also node 6 adjacent to the failure SF1 receives SF,6,7,L,ID and will send the corresponding signaling with ID code (SF,7,6,L,ID), reaching a stable scenario with "BS" signalings all over the ring.

[0081] As soon as also SF1 is cleared, the ring will reach the faultless stable condition, with the progressive removal of the "Bridge" and "Switch" actions from all the path termination nodes and the consequent signaling with "No Request, Idle" code (NR,5,6,S,ID and NR, 8,7,S,ID) issued by all the ring nodes, including nodes (6, 7) adjacent to the just cleared failure (SF2). See Figs. 9.7 and 9.8.

[0082] In view of the above detailed description, relating to some cases of single or double failure, the person skilled in the art can easily devise the actions that every node must perform in the event of a failure on other spans and/or in the case where more than two failures occur. Naturally, the present invention is applicable to all these cases and its scope covers all these cases and is limited only by the following claims.

[0083] As far as the practical realization is concerned, it will be understood that all the actions performed by every node or network element are the known Pass-Through, Bridge and Switch, squelching of any terminations of the Low-Priority channels involved in the protection and transmission of signaling, substantially of known type, actions. Therefore, the implementation of the present method does not require to change the physical structure of the existing network elements used in ring networks protected against possible failures. Any modifications must be carried out at level of consequent actions performed by the nodes affected by the protection mechanism, according to signalings already provided for and present in the standardized protocol and on the ground of ring map information, already provided for and processed, as well as traffic map that carries the allocation time-slot information, in every ring span of the single path that is installed.

[0084] Finally it is pointed out that, although the present invention has been described in detail with reference to SDH synchronous transmission, it applies, in similar manner, to other types of synchronous transmission, typically SONET. The fact that this type of signals has not been taken into account in the description shall not be interpreted as a limitation but merely as an example and in order to render the description clear. Therefore, for the purposes of the present description and of the annexed claims, the terminology used for SDH transmissions will include at least the corresponding SONET terminology and shall be read in this perspective.

Claims

1. Method of re-routing a path in a transoceanic MS shared protection ring network in the event of a failure on a span of said path, said ring network comprising network elements connected in a ring configuration by fiber spans, said fiber spans comprising high-priority (HP) channels and low-priority (LP) channels, said method comprising the step of performing a ring switch action by the MS shared protection mechanism, **characterized in that** a time slot interchange mechanism (TSI) is provided in said ring network and **in that** said method comprises the step of re-routing the path over the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the span affected by the failure.

2. Method according to claim 1, in which a further span of the path becomes affected by a failure, **characterized by** comprising the steps of: i) releasing the present re-routing that was performed because of the first failured span; ii) selecting one of the failured spans; and iii) re-routing the failured path over the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span that has been selected.

3. Method according to claim 1, in which a further span becomes affected by a failure, **characterized by** comprising the step of maintaining the re-routing action, performed because of the first failured span, should the persistency of the re-routing information be supported by the network elements of the ring network.

4. Method according to claim 2, **characterized in that** the step of selecting one of the failured spans comprises the step of considering the two spans adjacent to the switching nodes that are able to communicate with the termination nodes of path to be protected in the case where at least one further span of the path becomes affected by a failure.

5. Method according to any of claims 2 to 4, **characterized in that** the step of selecting one of the failured spans comprises the step of selecting the failured span adjacent to the switching node having higher or lower node identification ID.

6. Method according to any of claims 2 to 4, **characterized in that** the step of selecting one of the failured spans comprises the step of selecting the failured span adjacent to the switching node that comes first or last in the network ring map.

7. Method according to any of claims 2 to 4, **characterized in that** the step of selecting one of the failured spans comprises the step of selecting the failured span adjacent to the far west (W) or far east (E) switching node in the ring network.

8. Network element of a transoceanic MS shared protection ring network, said ring network comprising other network elements connected each other in a ring configuration by fiber spans, said fiber spans comprising high-priority (HP) channels and low-priority (LP) channels, said network element comprising means for performing ring switch actions, namely pass-through, bridge or switch actions, upon reception of corresponding signalings and means for generating and sending proper signalings in response to reception of corresponding signalings, a path being installed in said ring network, **characterized in that** a time slot interchange mechanism (TSI) is provided in said ring network and **in that** said network element comprises means for, in case of failure in a span of the installed path, re-routing the path over the time slot of the low priority (LP) channels corresponding to the time slot of the high priority (HP) channels of the failured span.

9. Network element according to claim 8, in which a further span of the path becomes affected by a failure, **characterized by** comprising: i) means for releasing the re-routing action performed because of the first failured span; ii) means for selecting one of the failured spans; and iii) means for re-routing the path over the time slot of the low priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span which has been selected.

10. Network element according to claim 8, in which a further span of the path becomes affected by a failure, **characterized by** comprising means for maintaining the re-routing action, performed because of the first failured span, should the persistency of the re-routing information be supported by the network elements of the ring network.

11. Network element according to claim 9, **characterized in that** said means for selecting one of the failured spans comprise means for considering the two spans adjacent to the switching nodes able to communicate with the termination nodes of path to be protected in the case where at least one further span of the path becomes affected by a failure.

12. Network element according to claim 9, said network element being a path termination node, **characterized by** comprising means for performing a Bridge&Switch action upon reception of two signalings comprising corresponding bridge requests with Bridge&Switch status code (BS) related to different spans.

13. Network element according to claim 9, said network element being a path non-termination node, **characterized by** comprising means for performing a pass-through action upon reception of at least one signaling comprising a bridge request with a Bridge&Switch status code (BS)

14. Network element according to claim 8 or 9, said network element being a path termination node, **characterized by** comprising means for performing a Bridge&Switch action upon reception of two signalings comprising corresponding bridge requests with Idle status code related to the same span.

15. Network element according to claim 9, said network element being a path non-termination node, **characterized by** comprising means for performing a pass-through action upon reception of at least one signaling comprising a bridge request with Idle status code.

16. Ring network comprising one or more network elements according to any of claims 8 to 15.

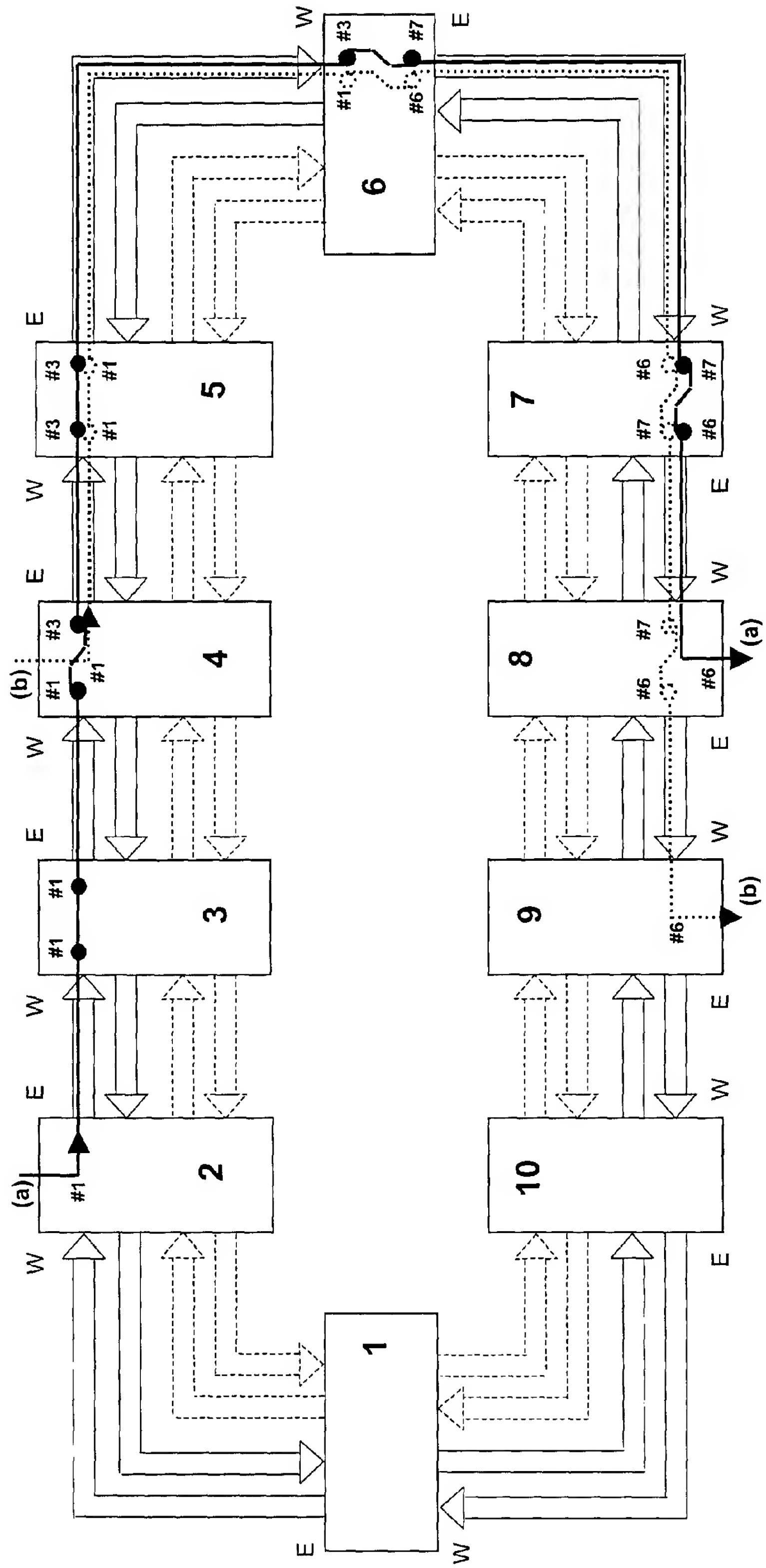


Fig. 1

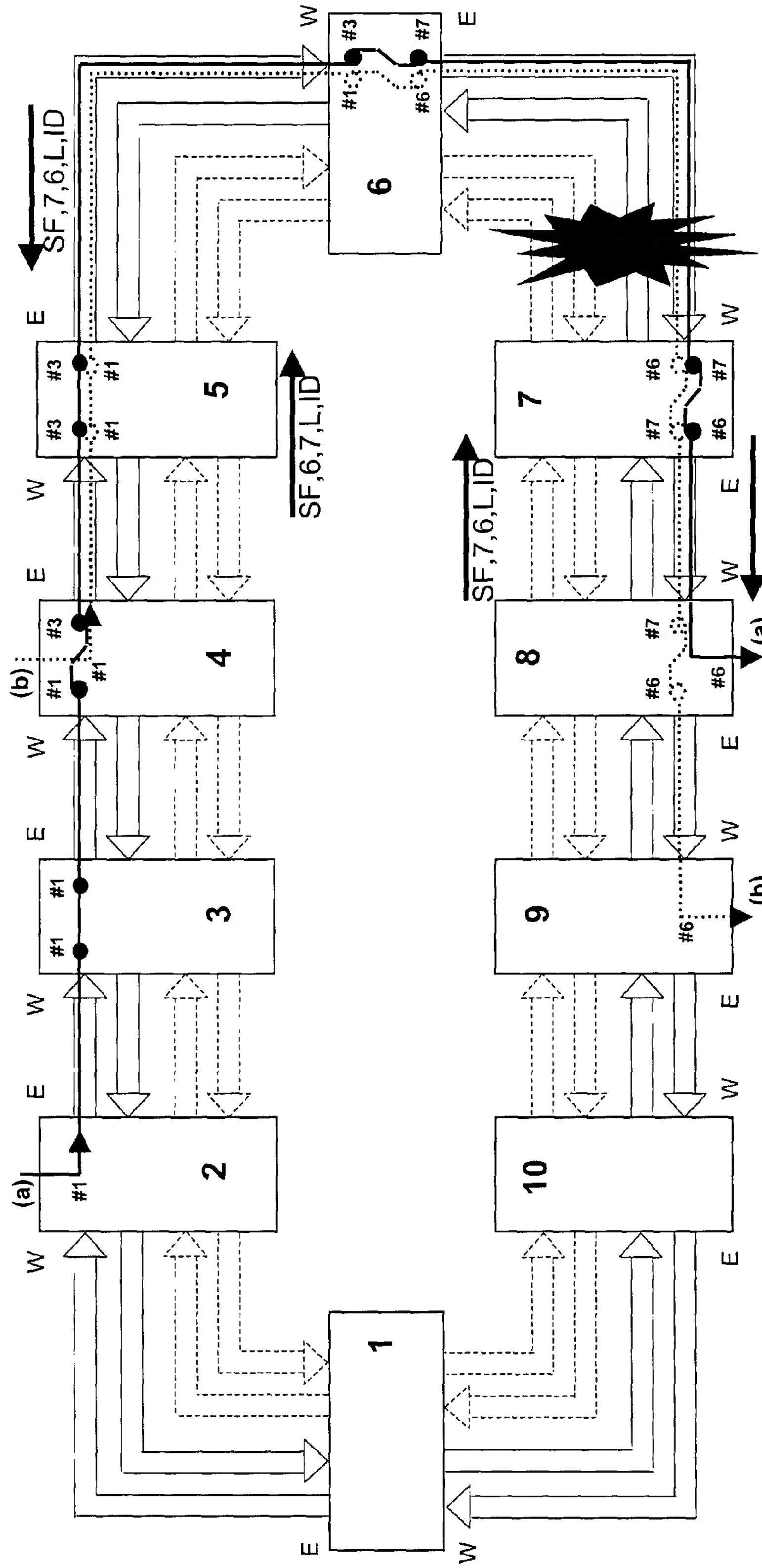


Fig. 2

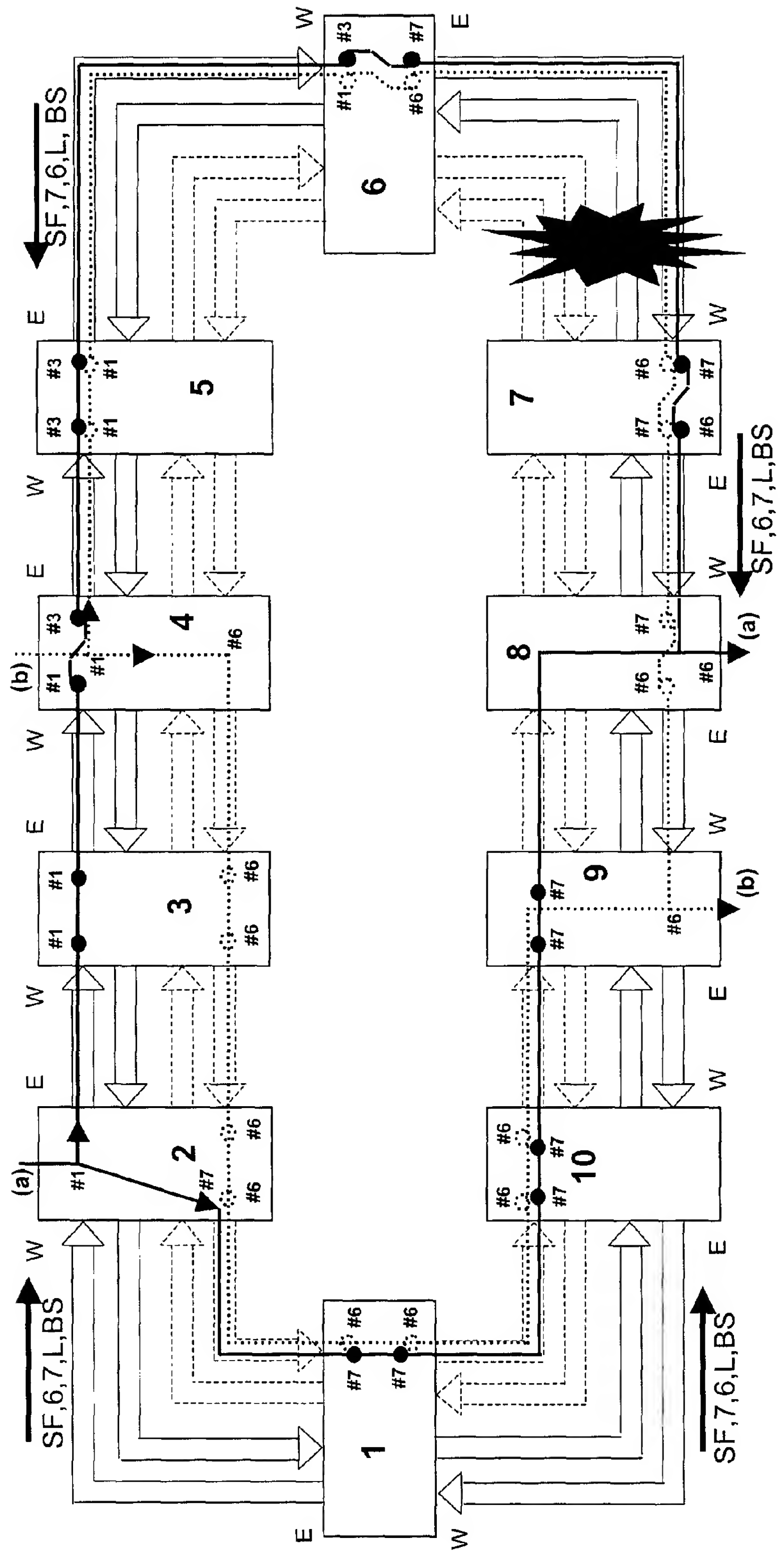


Fig. 3

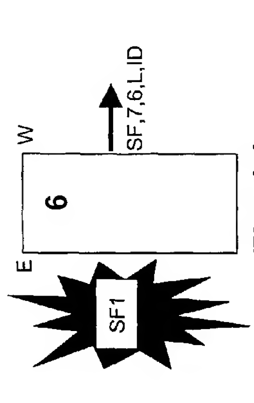


Fig. 4.1

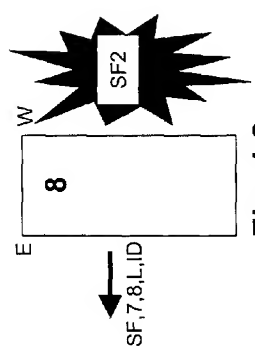


Fig. 4.2

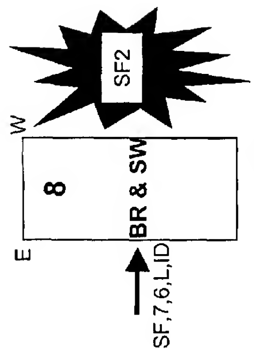
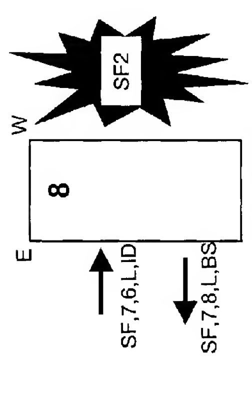


Fig. 4.3



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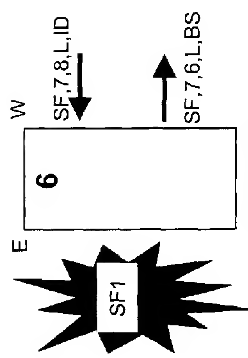

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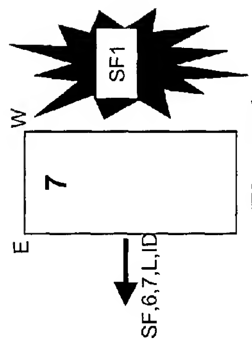

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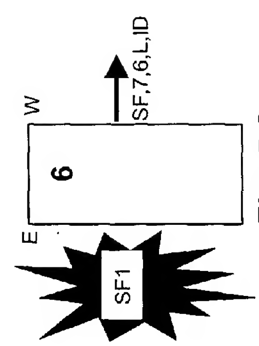

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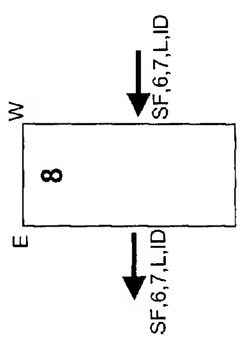

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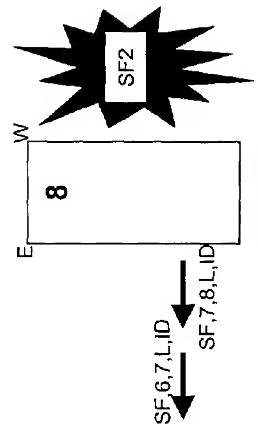

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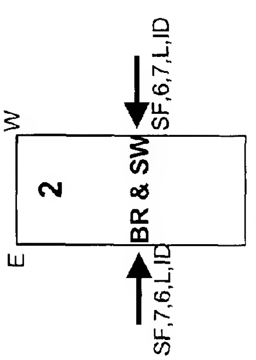

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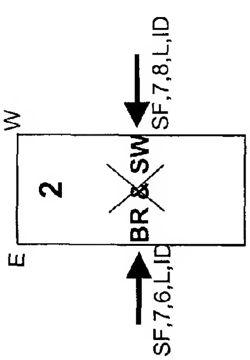

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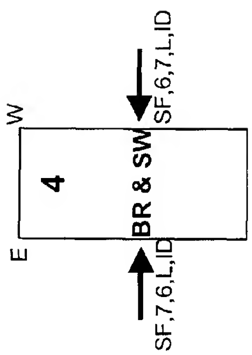

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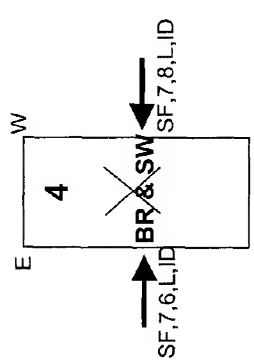

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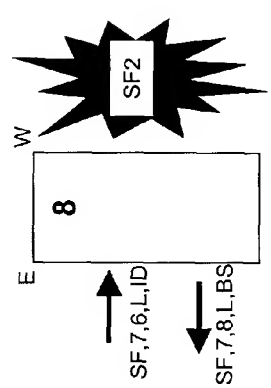

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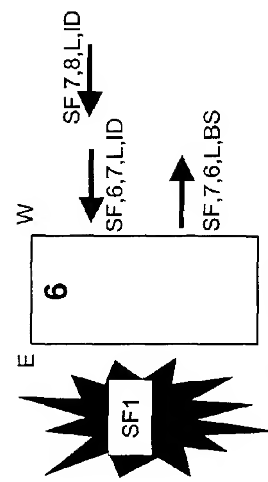

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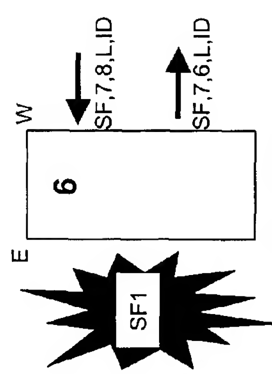

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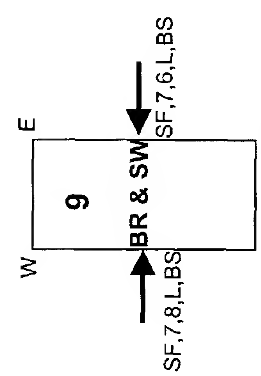

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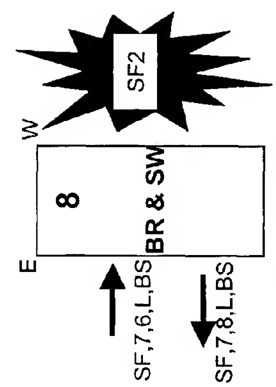

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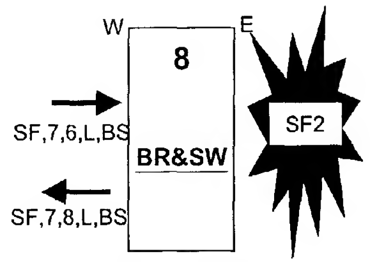

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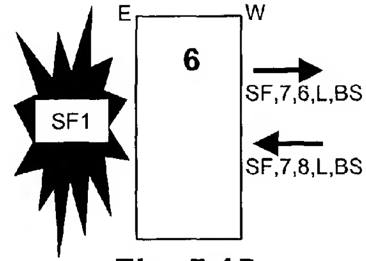

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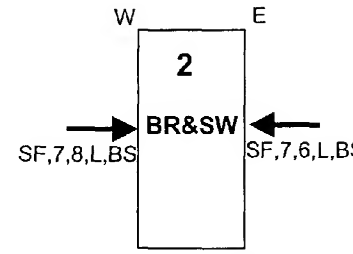

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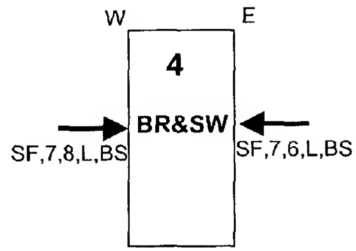

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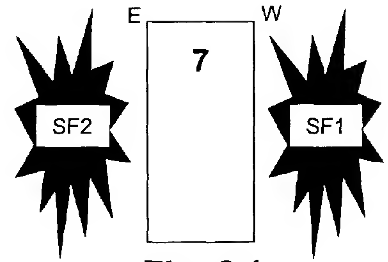

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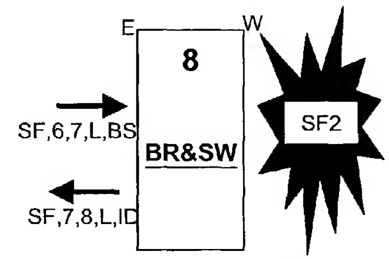

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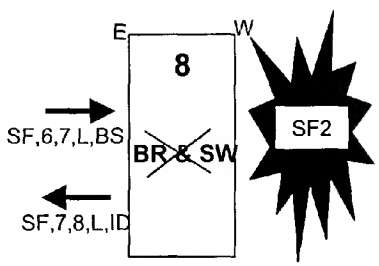

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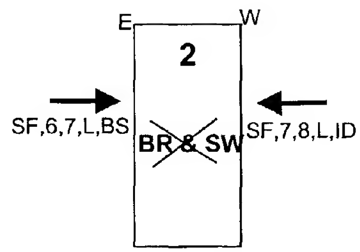

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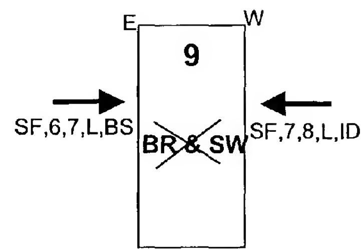

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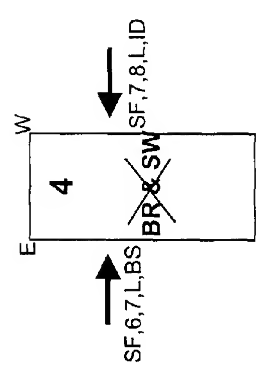

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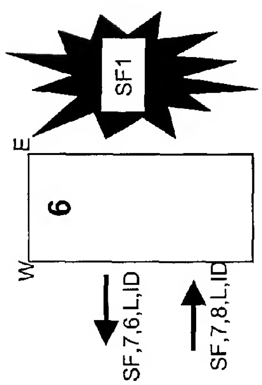

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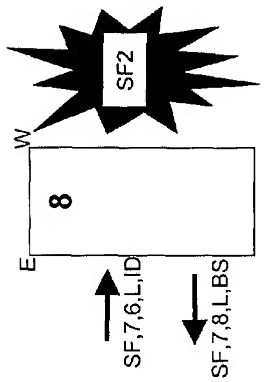

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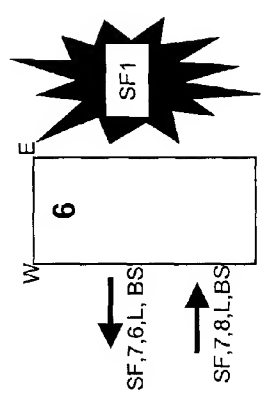

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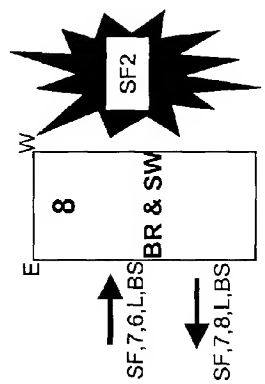

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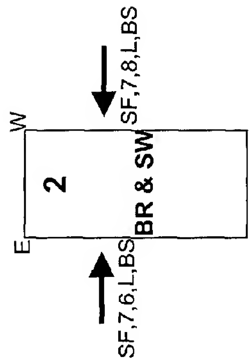

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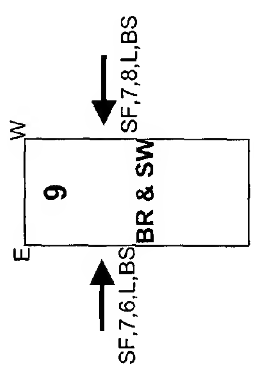

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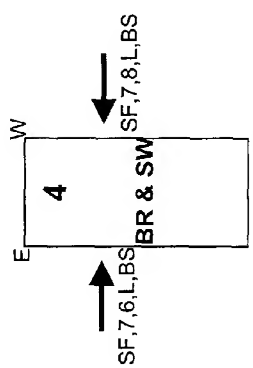

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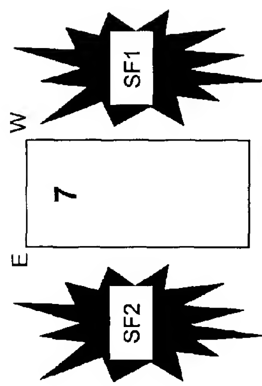

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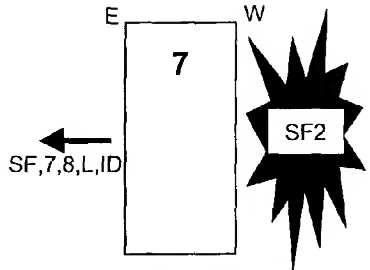

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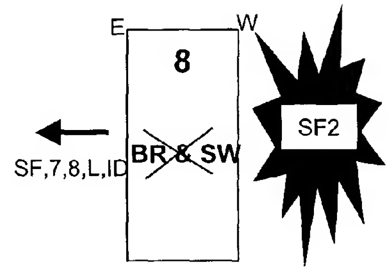

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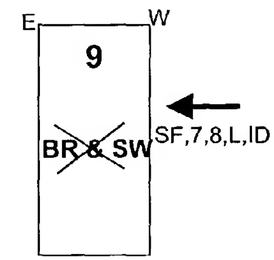

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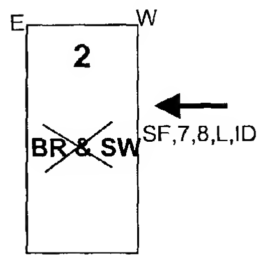

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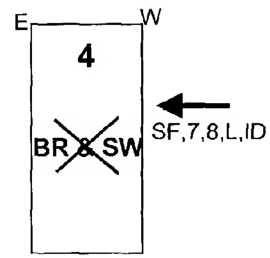


Fig. 7.6

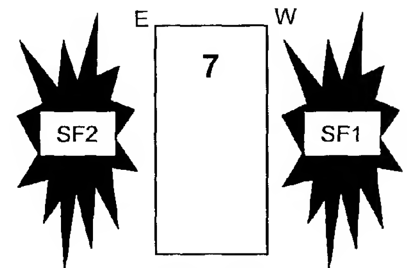
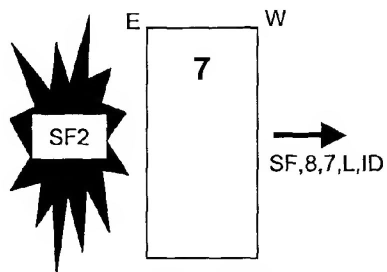


Fig. 8.1



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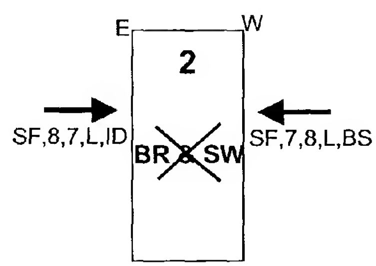

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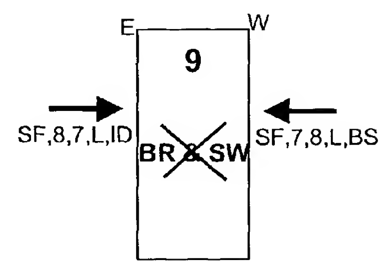

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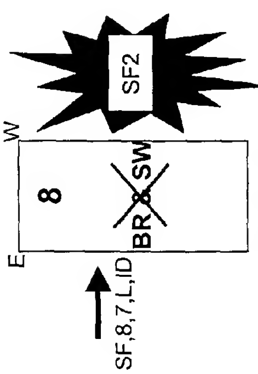

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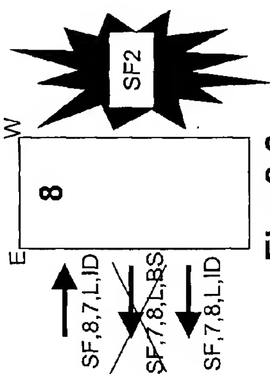

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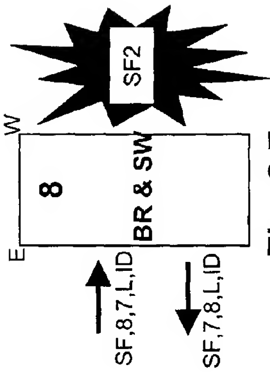

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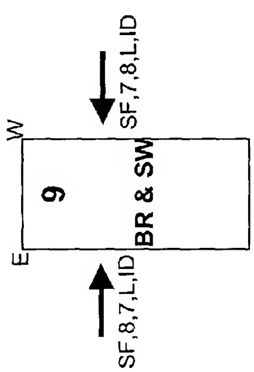

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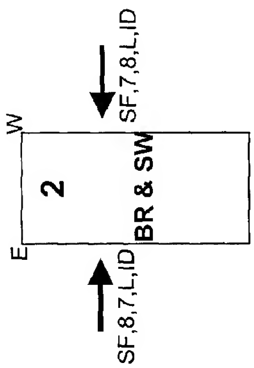
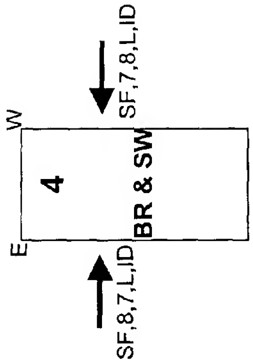


Fig.8.9



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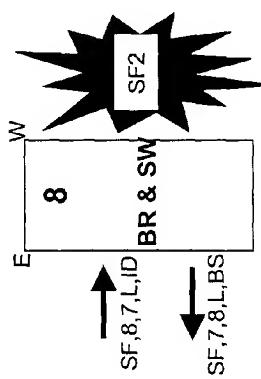

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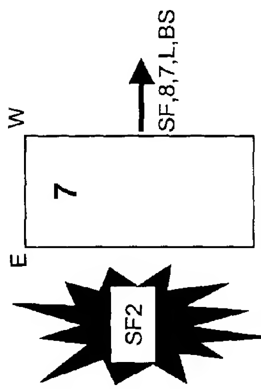

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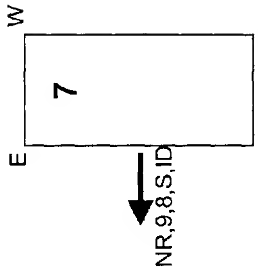

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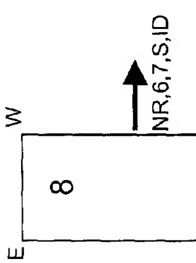

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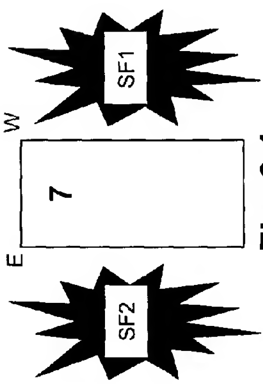

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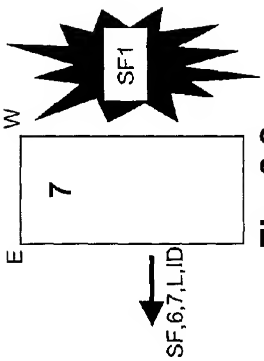

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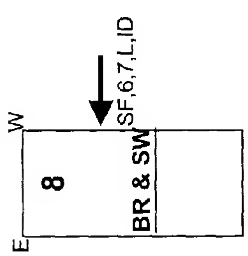

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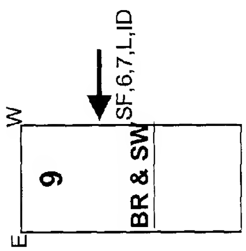

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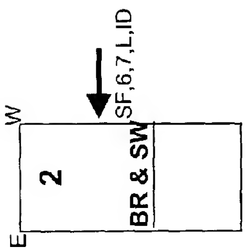

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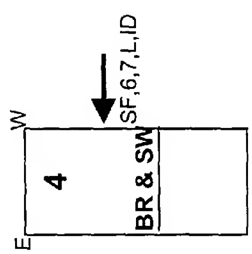

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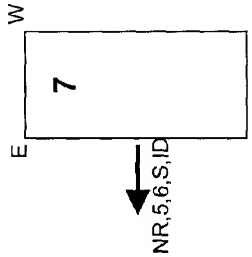

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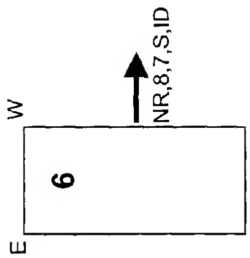

Fig. 9.8

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(54) Preamble design for multicarrier transmission over channels with multiple inputs and outputs

(57) One or more preambles are inserted into frames of Orthogonal Frequency Multiplexing (OFDM) -Multiple Input, Multiple Output (MIMO) signals. The preamble is received by the antennas of a receiver, decoded and compared to known values to provide synchronization, framing, channels estimation, offsets and other corrections to the transmitted signal.

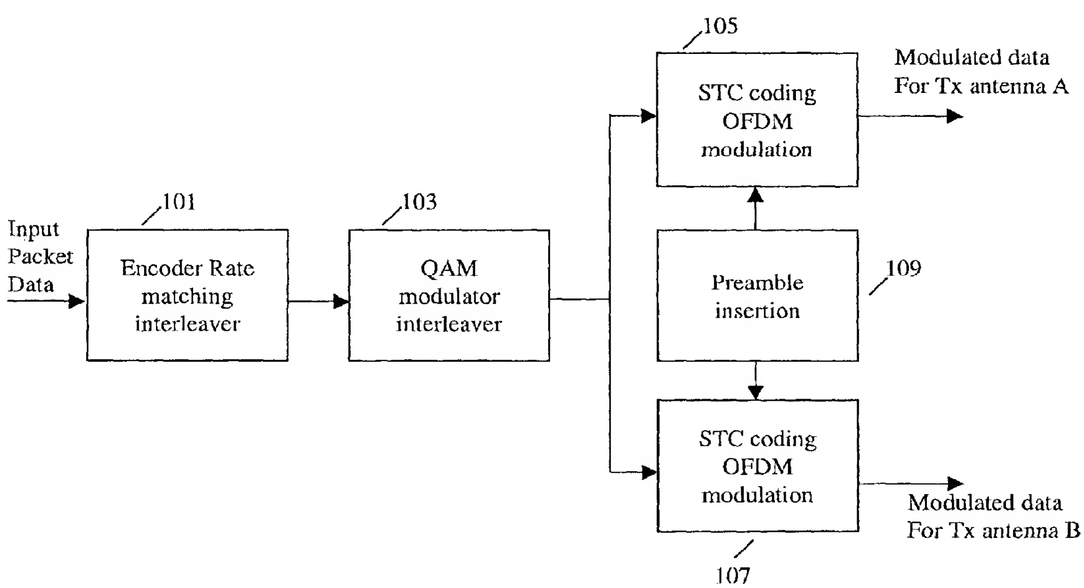

FIG. 1

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Description

FIELD OF THE INVENTION

[0001] The present invention is directed to the delivery of data via a wireless connection and, more particularly, to the accurate delivery of data at high rates via a wireless connection.

BACKGROUND OF THE INVENTION

[0002] The demand for services in which data is delivered via a wireless connection has grown in recent years and is expected to continue to grow. Included are applications in which data is delivered via cellular mobile telephony or other mobile telephony, personal communications systems (PCS) and digital or high definition television (HDTV). Though the demand for these services is growing, the channel bandwidth over which the data may be delivered is limited. Therefore, it is desirable to deliver data at high speeds over this limited bandwidth in an efficient, as well as cost effective, manner.

[0003] A known approach for efficiently delivering high speed data over a channel is by using Orthogonal Frequency Division Multiplexing (OFDM). The high-speed data signals are divided into tens or hundreds of lower speed signals that are transmitted in parallel over respective frequencies within a radio frequency (RF) signal that are known as sub-carrier frequencies ("sub-carriers"). The frequency spectra of the sub-carriers overlap so that the spacing between them is minimized. The sub-carriers are also orthogonal to each other so that they are statistically independent and do not create crosstalk or otherwise interfere with each other. As a result, the channel bandwidth is used much more efficiently than in conventional single carrier transmission schemes such as AM/FM (amplitude or frequency modulation), in which only one signal at a time is sent using only one radio frequency, or frequency division multiplexing (FDM), in which portions of the channel bandwidth are not used so that the sub-carrier frequencies are separated and isolated to avoid inter-carrier interference (ICI).

[0004] Further, each block of data is converted into parallel form and mapped into each subcarrier as frequency domain symbols. To get time domain signals for transmission, an inverse discrete Fourier transform or its fast version, IFFT, is applied to the symbols. The symbol duration is much longer than the length of the channel impulse response so that inter-symbol interference is avoided by inserting a cyclic prefix for each OFDM symbol. Thus, OFDM is much less susceptible to data loss caused by multipath fading than other known techniques for data transmission. Also, the coding of data onto the OFDM sub-carriers takes advantage of frequency diversity to mitigate loss from frequency-selective fading when forward error correction (FEC) is applied.

[0005] In addition to having greater spectral efficiency, i.e. more bps/Hz, than conventional transmission schemes, the OFDM spectral efficiency is further enhanced because the spectrum can be made to look like a rectangular window so that all frequencies are similarly utilized. Moreover, OFDM is less sensitive to timing errors because the timing errors are translated to a phase offset in the frequency domain.

[0006] Another approach to providing more efficient use of the channel bandwidth is to transmit the data using a base station having multiple antennas and then receive the transmitted data using a remote station having multiple receiving antennas, referred to as Multiple Input-Multiple Output (MIMO). The data may be transmitted such there is spatial diversity between the signals transmitted by the respective antennas, thereby increasing the data capacity by increasing the number of antennas. Alternatively, the data is transmitted such that there is temporal diversity between the signals transmitted by the respective antennas, thereby reducing signal fading.

[0007] Presently, MIMO systems either are designed to transmit signals having spatial diversity or are designed to transmit signals having temporal diversity. It is therefore desirable to provide a common system that can deliver signals with either spatial diversity or temporal diversity depending on the transmission environment.

[0008] It is further desirable to provide a system that has the advantages of both an OFDM system as well as those of a MIMO system. Such a system would transmit the OFDM symbols over a plurality of channels with either spatial diversity or temporal diversity between the symbols. However, when the signals are received at the remote station, the framing and timing of the received signals and the frequency and sampling clock offsets must be determined so that the information contained in the received signals may be recovered. Further, the signals may be distorted because of transmitter imperfections as well as because of environmental effects and interference which change the frequencies of the channels and may also increase the bit error rate (BER). Additionally, the gain of the received signals must be controlled.

[0009] Accordingly, it is advantageous to provide a system that can efficiently transfer data from a transmitter to a receiver over multiple channels.

SUMMARY OF THE INVENTION

[0010] The present invention provides a preamble that is inserted into a signal frame and which corresponds to a respective transmitter antenna. The preamble is matched to known values by a respective receiver to decode the signals and permit multiple signals to be transferred from the transmitter to the receiver.

[0011] In accordance with an aspect of the invention, a preamble portion of a data signal is configured for

transmission over a plurality of sub-carriers by at least two antennas of a transmitter device. A respective pseudo-noise (PN) code is assigned to each of the at least two antennas. Each of the plurality of sub-carriers is assigned to a respective one of the at least two antennas. Each of the plurality of sub-carriers is modulated as a function of the respective pseudo-noise (PN) code that is assigned to a same one of the at least two antennas as the each of the plurality of sub-carriers such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of the at least two antennas. Each of the plurality of modulated sub-carriers is delivered to its assigned transmitter. Each the plurality of modulated sub-carriers using its assigned transmitter is transmitted at substantially a same time.

[0012] According to another aspect of the invention, a preamble portion of a data signal is configured for transmission over a plurality of sub-carriers by at least two transmitter devices each having at least two antennas. A respective pseudo-noise (PN) code is assigned to each of the at least two antennas. Each of the plurality of sub-carriers is assigned to a respective one of the at least two transmitter devices. Each of the plurality of sub-carriers is modulated as a function of the respective pseudo-noise (PN) code that is assigned to a same one of the at least two transmitter devices to which the each of the plurality of sub-carriers is assigned such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of the at least transmitter devices. Each of the plurality of modulated sub-carriers using each of the at least two antennas of its assigned transmitter device at substantially a same time.

[0013] Other features and advantages of the present invention will become apparent from the following detailed description of the invention with reference to the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

[0014] The invention will now be described in greater detail in the following detailed description with reference to the drawings in which:

Figure 1 is a block diagram showing an example of transmitter arrangement for generating OFDM-MIMO signals that include preambles according to the invention.

Figure 2 is a diagram showing an example of a frame, slot and symbol structure for signals of the invention.

Figure 3 is a diagram showing an arrangement of preamble, pilot and data symbols according to subcarrier frequency and time in accordance with the invention.

Figure 4 is a diagram showing an example of an arrangement in the frequency domain of pilot carriers transmitted by a pair of transmitter antennas according to the invention.

Figure 5 is a diagram showing an example of an arrangement in the frequency domain of pilot carriers transmitted by plural base stations according to the invention.

Figure 6 is a diagram showing an arrangement of preambles and data within one or more frames in accordance with the invention.

Figure 7 is a block diagram showing an example of a receiver arrangement for receiving and decoding OFDM-MIMO signals that include preambles according to the invention.

DETAILED DESCRIPTION OF THE INVENTION

[0015] The present invention provides an Orthogonal Frequency Divisional Multiplexing (OFDM) signal that is delivered using Multiple-Input and Multiple-Output (MIMO) transmitter and receiver antennas. In a conventional OFDM system, a block of data is represented by a frequency domain signal vector S that may be comprised of real or complex values. The vector S may be comprised of, for example, 1024 elements, namely S = (s0, s1, ..., s1024). Each element of the frequency domain signal vector S is used to modulate a respective sub-carrier frequency of the carrier signal to obtain OFDM symbols. The frequency domain signal vector S is then converted into the time domain, such as using a inverse fast Fourier transform (IFFT), to obtain a time domain vector V = IFFT(S) = (v0, v1, ..., v1024). A cyclic prefix, comprised of the last elements of the vector V, is then inserted at the front of the vector V to obtain vector V' = (v1000, v1001, ..., v1024, v0, v1, ..., v1024). The elements of the vector V' are then transmitted serially by a single transmitter antenna over the single channel to a receiver having a single receiver antenna.

[0016] In an OFDM-MIMO system of the invention, by contrast, the OFDM symbols are transmitted in the time domain using multiple antennas to concurrently transmit the symbols over the same sub-carriers to multiple receiver antennas. However, when the signals are detected by the multiple antenna, they may be distorted and must also be synchronized and framed properly to avoid errors.

[0017] Thus, the invention provides one or more preambles which are is inserted between the OFDM data symbols within OFDM frames in the time domain. The preamble includes training symbols which include a training sequence for different antennas, also known as pilot symbols.

[0018] Figure 1 shows an arrangement an OFDM transmitter employed by the invention. An encoder and rate matching interleaver 101 receives a stream of data bits and divides the stream of data bits into segments of B bits each, such as segments of 1024 bits. A block and/or a convolutional coding scheme is then carried out on the segments of B bits to introduce error correcting and/or error-detecting redundancy. The segments of B bits

are then respectively subdivided into 2N sub-segments of m bits each, where m typically has a value of from two to six.

[0019] The encoder and rate matching interleaver 101 then delivers the sub-segments of data to a quadrature amplitude modulation (QAM) modulator and interleaver 103 which maps the sub-segments onto corresponding complex-valued points in a 2^m-ary constellation. A corresponding complex-valued 2^m-ary QAM sub-symbol, $c_k=a_k+jb_k$, that represent a discrete value of phase and amplitude, where -N # k # N, is assigned to represent each of the sub-segments such that a sequence of frequency-domain sub-symbols is generated. The QAM modulator and interleaver 103 also assigns the value $c_o=0$ to the zero-frequency sub-carrier and interleaves any additional zeroes that may be required for later interpolation into the sequence of frequency-domain sub-symbols.

[0020] The QAM modulator and interleaver 103 then delivers the sequence of frequency-domain sub-symbols to one of space time coding (STC) and OFDM modulation circuits 105 and 107 which employs an inverse fast Fourier transform (IFFT) to modulate the phase and amplitude of the sub-carriers and also space time code the sub-carriers to incorporate either spatial diversity or temporal diversity between the sub-carriers. Each of the complex-valued, frequency-domain sub-symbols c_k is used to modulate the phase and amplitude of a corresponding one of 2N+1 sub-carrier frequencies over a symbol interval T_s. The sub-carriers are each represented by change value e^{-2Jjf_k} and have baseband frequencies of $f_k=k/T_s$, where k is the frequency and is an integer in the range $-N \le k \le N$. A plurality of digital time-domain OFDM symbols of duration T_s are thus generated according to the relation:

$$u(t)=\sum_{k=-N}^{N} c_k \exp(-2\pi jf^k),$$

where $0 \le t \le T_s$.

[0021] The modulated sub-carriers are each modulated according to a sinc x=(sin x)/x function in the frequency domain, with a spacing of $1/T_s$ between the primary peaks of the sub-carriers, so that the primary peak of a respective sub-carrier coincides with a null the adjacent sub-carriers. Thus, the modulated sub-carriers are orthogonal to one another though their spectra overlap.

[0022] A preamble insertion circuit 109 stores and periodically inserts at least one preamble into the modulated sub-carriers respectively generated by the STC and OFDM modulation circuits 105 and 107 according to the invention. The STC and OFDM modulation circuits 105 and 107 then deliver the modulated sub-carriers and the preambles in the time domain to their respective antennas (not shown) for transmission.

[0023] Preferably, the pilot symbols of the preamble are initially generated in the frequency domain by modulating frequency domain sub-carriers using a psuedo-noise (PN) code that is unique to each transmitter antenna. Then, the frequency domain pilot symbol sequence is converted to the time domain using an inverse fast Fourier transform (IFFT). The time domain pilot symbols are then stored in a memory in the pilot insertion circuit 109 and are then periodically inserted into the time domain OFDM-MIMO signal, such as at the beginning of a frame.

[0024] Figure 2 shows an example of a structure of the transmitted OFDM-MIMO signal in the time domain. The signal is formatted as a plurality of frames 201, Each frame includes plural slots 203. The first slot of each frame includes a preamble that is located at the beginning of the slot. The preamble includes two training symbols 205 and plural symbols 207.

[0025] Figure 3 shows the transmitted symbols arranged according to increasing time and increasing sub-carrier frequency. In the time domain, the first two symbols of a frame are preamble symbols, as described above. Thereafter, data symbols or pilot symbols are transmitted, depending on the sub-carrier frequency, until the next preamble symbols are transmitted.

[0026] Figure 4 illustrates, in greater detail, an example of a preamble shown in Figure 3, referred to as Preamble 1. The preamble is broadcast by a single base station having at least two transmitter antennas. Each transmitter antenna transmits respective pairs of identical training symbols, also known as pilot carrier symbols, at a given sub-carrier frequency.

[0027] The sub-carrier frequencies are divided into groups which are each assigned to a respective transmitter antenna. For example, Figure 3 shows two transmitter antennas where, for example, the even numbered sub-carriers are assigned to Antenna Tx1 and the odd numbered sub-carriers are assigned to Antenna Tx2. The pilot symbols for each antenna are orthogonal in the frequency domain in an interlaced transmission patterns, and the pilot symbols are superimposed in the time domain.

[0028] A unique psuedo-noise (PN) code is assigned to each transmitter antenna to define the pilot symbols used to modulate the sub-carrier frequencies. The values of the pilot symbols that are transmitted are known to the receiver and may be used by the receiver to determine the framing of the transmitted signal, to determine the timing of the transmitted signal, to estimate the frequency and timing clock offsets of the receiver, to estimate the distortion in the transmitted sub-carrier channels, and to estimate the carrier-to-interference (C/I) ratio of the transmitted signals.

[0029] When more than one base transceiver station (BTS) transmits to a receiver, another example of a preamble, referred to as Preamble 2, may be used, such

as for fast cell switching applications. Two training symbols are used. However, the sub-carrier frequencies are divided among the BTSs, and each BTS uses a respective PN code to modulate its assigned sub-carrier frequencies. The sub-carrier frequencies assigned to particular BTS, as well as the PN code assigned to each BTS, are known to the receiver and may be used to provide co-channel interference cancellation between the BTSs, may be used to estimate the sub-carrier channels used by adjacent BTSs, and may be used as pilot symbols to track the sub-carrier channels.

[0030] Figure 5 illustrates and example of Preamble 2 in which 6 BTSs each transmit using two respective antennas. The two training symbols are assigned to Antenna Tx1 and Antenna Tx2, respectively. The sub-carrier frequencies are divided into 6 groups, each as signed to a BTSs.

[0031] Figure 6 illustrates an example of the insertion of preambles into one or more frames 601. A preamble, shown here as Preamble 1, is inserted at the beginning of the frame. Then, depending on the length of the frame and the channel conditions, additional preambles may be inserted at locations within the frame. As an example, Preamble 2 is inserted in the middle of the frame. A further preamble may be inserted at the end of the frame or at the beginning of the next frame.

[0032] Figure 7 is an example of a receiver that receives and decodes the preambles of the present invention. The receiver may be an Internet network terminal, cellular or wireless telephone, or other device that is able to receive OFDM-MIMO signals.

[0033] RF signals received by receiver antennas A and B (not shown) are delivered to respective circuits 701, 703 which convert the analog OFDM signals into digital signals and use the preamble of the signal to synchronize the signal and determine the frame boundaries of the transmitted data, such as by using sliding correlation. The framed data is then converted into vector form.

[0034] To obtain better framing synchronization, a fine synchronization stage is used by checking the correlation between received signals with known signals that are stored in the OFDM receiver memory. The orthogonal property of PN pilots in the training symbols is utilized to separate MIMO channels and perform the fine synchronization. The synchronization may be performed either in the time domain or in the frequency domain. The MIMO system makes the fine synchronization more robust to multi-path fading due to the separation of the MIMO channel correlators.

[0035] The synchronization operation is described in greater detail in U.S. Application No. 09/751,881, titled "Synchronization in a Multiple-input/multiple-output (MIMO) Orthogonal Frequency Division Multiplexing (OFDM) System For Wireless Applications", filed December 29, 2000 by the applicants of the present application, and incorporated herein by reference.

[0036] After timing synchronization, the FFT window is determined, and the received OFDM signals are framed and are transferred into the frequency domain. The pilot channel can be used to estimate the frequency and sampling clock. The performance can be improved by averaging the results obtained from the different MIMO channels.

[0037] The digital signals are also corrected for any differences between the oscillation frequency of the local oscillator of the transmitter system and the oscillation frequency of the local oscillator of the receiver system. A correction signal is used in generating the data vectors.

[0038] The circuits 701 and 703 then deliver the data vectors to their demodulators 705, 707 which removes unneeded cyclical extensions in the data vector and performs a discrete Fourier transform (DFT) or a Fast Fourier Transform (FFT) that demodulates the data vectors to recover the original sequences of frequency domain sub-symbols. The demodulators 705, 707 then deliver the frequency domain sub-symbols to a STC decoder 713 which decodes the sub-symbols. The STC decoder also uses the preamble portion of the sub-symbols to correct for co-channel interference.

[0039] The demodulators 705, 707 also deliver the preamble portion of the frequency domain sub-symbols to their respective channel estimators 709, 711 which use the detected sub-symbols and the known values of the sub-symbols to estimate the values of channel responses vectors that are delivered to the STC decoder 713 to compensate for distortions in the received signal.

[0040] The STC decoder 713 then delivers the decoded sub-symbols to circuit 715 which performs a QAM demodulation and de-interleaving, and further decodes the sub-symbols to obtain the original raw bit stream.

[0041] The operation of the MIMO-OFDM receiver system and the channel estimation are described in greater detail in U.S. Application Nos. 09/750,804 titled "Adaptive Time Diversity and Spatial Diversity for OFDM and 09/751,166 titled "Channel Estimation for a MIMO OFDM System", both filed December 29, 2000 and incorporated herein by reference.

[0042] Although the present invention has been described in relation to particular embodiments thereof, many other variations and modifications and other uses may become apparent to those skilled in the art. It is preferred, therefore, that the present invention be limited not by this specific disclosure herein, but only by the appended claims.

Claims

1. A method of configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two antennas of a transmitter device, said method comprising:

   assigning a respective pseudo-noise (PN) code

to each of said at least two antennas;
assigning each of said plurality of sub-carriers to a respective one of said at least two antennas;
modulating each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two antennas as said each of said plurality of sub-carriers such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least two antennas;
delivering each of said plurality of modulated sub-carriers to its assigned transmitter; and
transmitting, at substantially a same time, each said plurality of modulated sub-carriers using its assigned transmitter.

2. The method of claim 1 wherein said data signal comprises an Orthogonal Frequency Division Multiplexing (OFDM) signal.

3. The method of claim 1 wherein said data signal is comprised of a plurality of frames, each of said frames being comprised of a plurality of time slots, each of said time slots including a plurality of symbols, and said method further comprises: inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a first one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned antenna.

4. The method of claim 3 further comprising: inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a further one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned antenna.

5. A method of configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two transmitter devices each having at least two antennas, said method comprising:

assigning a respective pseudo-noise (PN) code to each of said at least two antennas;
assigning each of said plurality of sub-carriers to a respective one of said at least two transmitter devices;
modulating each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two transmitter devices to which said each of said plurality of sub-carriers is assigned such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least transmitter devices; and
transmitting, at substantially a same time, each of said plurality of modulated sub-carriers using each of said at least two antennas of its assigned transmitter device.

6. The method of claim 5 wherein said signal comprises an Orthogonal Frequency Division Multiplexing (OFDM) signal.

7. The method of claim 5 wherein said data signal is comprised of a plurality of frames, each of said frames being comprised of a plurality of time slots, each of said time slots including a plurality of symbols, and said method further comprises: inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a first one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned transmitter device.

8. The method of claim 7 further comprising: inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a further one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned transmitter device.

9. An apparatus for configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two antennas of a transmitter device, said apparatus comprising:

a preamble insertion circuit configured to:

assign a respective pseudo-noise (PN) code to each of said at least two antennas;
assign each of said plurality of sub-carriers to a respective one of said at least two antennas; and
modulate each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two antennas as said each of said plurality of sub-carriers such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least two antennas; and
a coding circuit configured to deliver each of said plurality of modulated sub-carriers to its assigned transmitter; said transmitter antenna being configured to transmit, at substantially a same time, each said plurality of modulated sub-carriers using its assigned transmitter.

10. The apparatus of claim 9 wherein said data signal

comprises an Orthogonal Frequency Division Multiplexing (OFDM) signal.

11. The apparatus of claim 9 wherein said data signal is comprised of a plurality of frames, each of said frames being comprised of a plurality of time slots, each of said time slots including a plurality of symbols, and wherein said coding circuit is further configured to: insert said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a first one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned antenna.

12. The method of claim 11 wherein said coding circuit is further configured to: insert said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a further one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned antenna.

13. An apparatus for configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two transmitter devices each having at least two antennas, said apparatus comprising:

a preamble insertion circuit configured to:

assign a respective pseudo-noise (PN) code to each of said at least two antennas;

assign each of said plurality of sub-carriers to a respective one of said at least two transmitter devices; and

modulate each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two transmitter devices to which said each of said plurality of sub-carriers is assigned such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least transmitter devices; said at least two antennas of said at least two transmitter devices being configured to transmit, at substantially a same time, each of said plurality of modulated sub-carriers using each of said at least two antennas of its assigned transmitter device.

14. The apparatus of claim 13 wherein said signal comprises an Orthogonal Frequency Division Multiplexing (OFDM) signal.

15. The apparatus of claim 13 wherein said data signal is comprised of a plurality of frames, each of said frames being comprised of a plurality of time slots, each of said time slots including a plurality of symbols, and wherein said coding circuit is further configured to: insert said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a first one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned transmitter device.

16. The apparatus of claim 15 wherein said coding circuit is further configured to: insert said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a further one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned transmitter device.

17. An apparatus for configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two antennas of a transmitter device, said apparatus comprising:

means for assigning a respective pseudo-noise (PN) code to each of said at least two antennas;
means for assigning each of said plurality of sub-carriers to a respective one of said at least two antennas;
means for modulating each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two antennas as said each of said plurality of sub-carriers such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least two antennas;
means for delivering each of said plurality of modulated sub-carriers to its assigned transmitter; and
means for transmitting, at substantially a same time, each said plurality of modulated sub-carriers using its assigned transmitter.

18. An apparatus for configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two transmitter devices each having at least two antennas, said apparatus comprising:

means for assigning a respective pseudo-noise (PN) code to each of said at least two antennas;
means for assigning each of said plurality of sub-carriers to a respective one of said at least two transmitter devices;
means for modulating each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a

same one of said at least two transmitter devices to which said each of said plurality of sub-carriers is assigned such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least transmitter devices; and
means for transmitting, at substantially a same time, each of said plurality of modulated sub-carriers using each of said at least two antennas of its assigned transmitter device.

19. A readable medium comprising:

instructions for configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two antennas of a transmitter device, said instructions comprising:

instructions for assigning a respective pseudo-noise (PN) code to each of said at least two antennas;
instructions for assigning each of said plurality of sub-carriers to a respective one of said at least two antennas;
instructions for modulating each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two antennas as said each of said plurality of sub-carriers such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least two antennas;
instructions for delivering each of said plurality of modulated sub-carriers to its assigned transmitter; and
instructions for transmitting, at substantially a same time, each said plurality of modulated sub-carriers using its assigned transmitter.

20. The medium of claim 19 wherein said data signal comprises an Orthogonal Frequency Division Multiplexing (OFDM) signal.

21. The medium of claim 19 wherein said data signal is comprised of a plurality of frames, each of said frames being comprised of a plurality of time slots, each of said time slots including a plurality of symbols, and further comprising: instructions for inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a first one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned antenna.

22. The medium of claim 21 further comprising: instructions for inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a further one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned antenna.

23. A readable medium comprising:

instructions for configuring a preamble portion of a data signal for transmission over a plurality of sub-carriers by at least two transmitter devices each having at least two antennas, said instructions for comprising:

instructions for assigning a respective pseudo-noise (PN) code to each of said at least two antennas;
instructions for assigning each of said plurality of sub-carriers to a respective one of said at least two transmitter devices;
instructions for modulating each of said plurality of sub-carriers as a function of said respective pseudo-noise (PN) code that is assigned to a same one of said at least two transmitter devices to which said each of said plurality of sub-carriers is assigned such that a plurality of modulated sub-carriers are obtained that are each assigned to a respective one of said at least transmitter devices; and
instructions for transmitting, at substantially a same time, each of said plurality of modulated sub-carriers using each of said at least two antennas of its assigned transmitter device.

24. The medium of claim 23 wherein said signal comprises an Orthogonal Frequency Division Multiplexing (OFDM) signal.

25. The medium of claim 23 wherein said data signal is comprised of a plurality of frames, each of said frames being comprised of a plurality of time slots, each of said time slots including a plurality of symbols, and further comprising: instructions for inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a first one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned transmitter device.

26. The medium of claim 25 further comprising: instructions for inserting said each of said plurality of modulated sub-carriers into at least a first two of said plurality of symbols within a further one of said plurality of time slots prior to delivering said plurality of modulated sub-carriers to its assigned transmitter device.

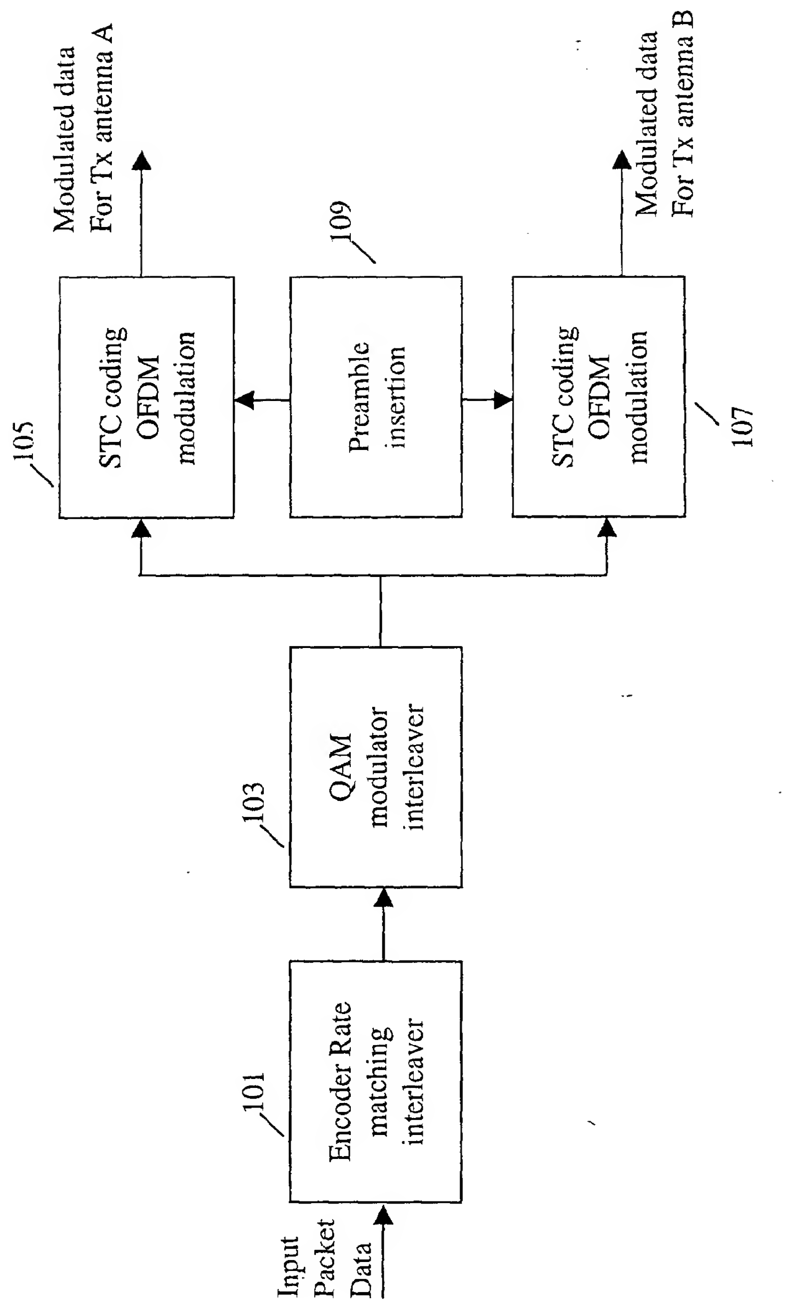
Input Packet Data
101
Encoder Rate matching interleaver
103
QAM modulator interleaver
105
STC coding OFDM modulation
Modulated data For Tx antenna A
109
Preamble insertion
107
STC coding OFDM modulation
Modulated data For Tx antenna B

FIG. 1

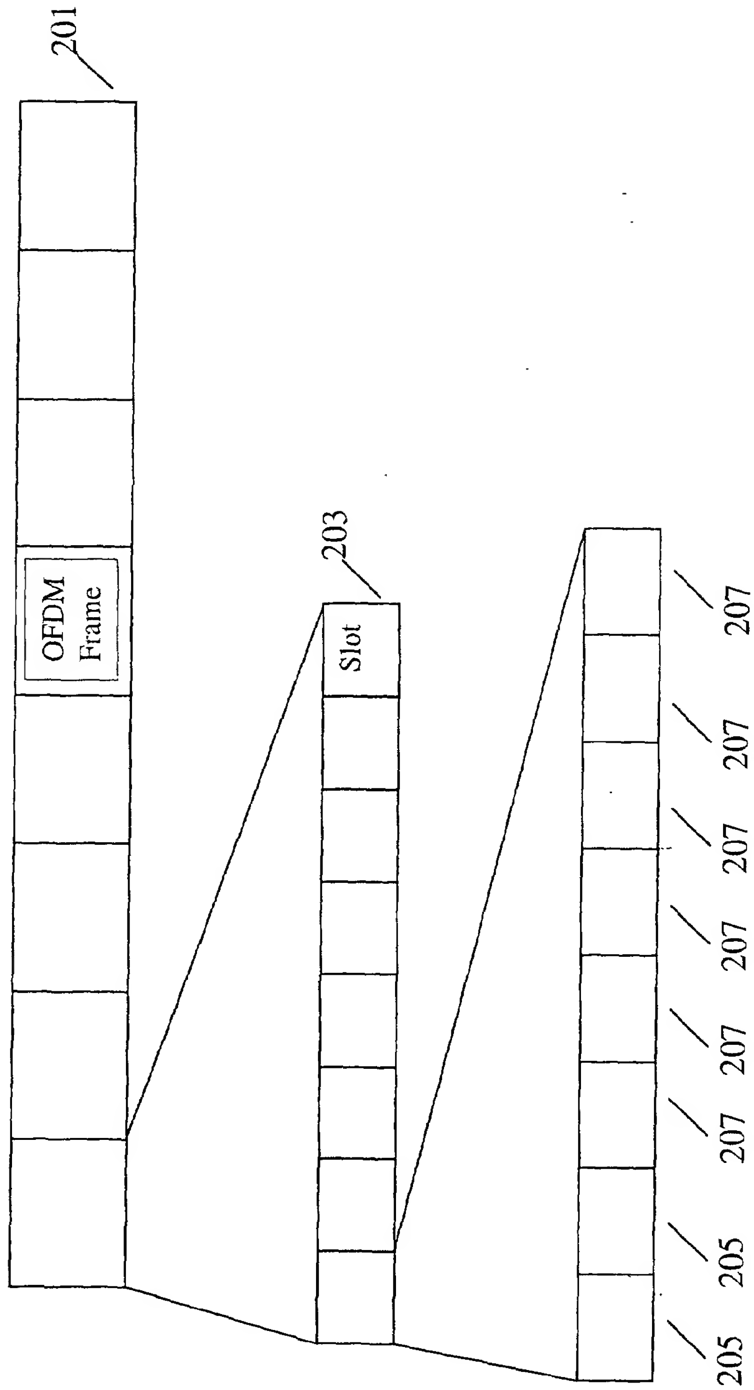

FIG. 2

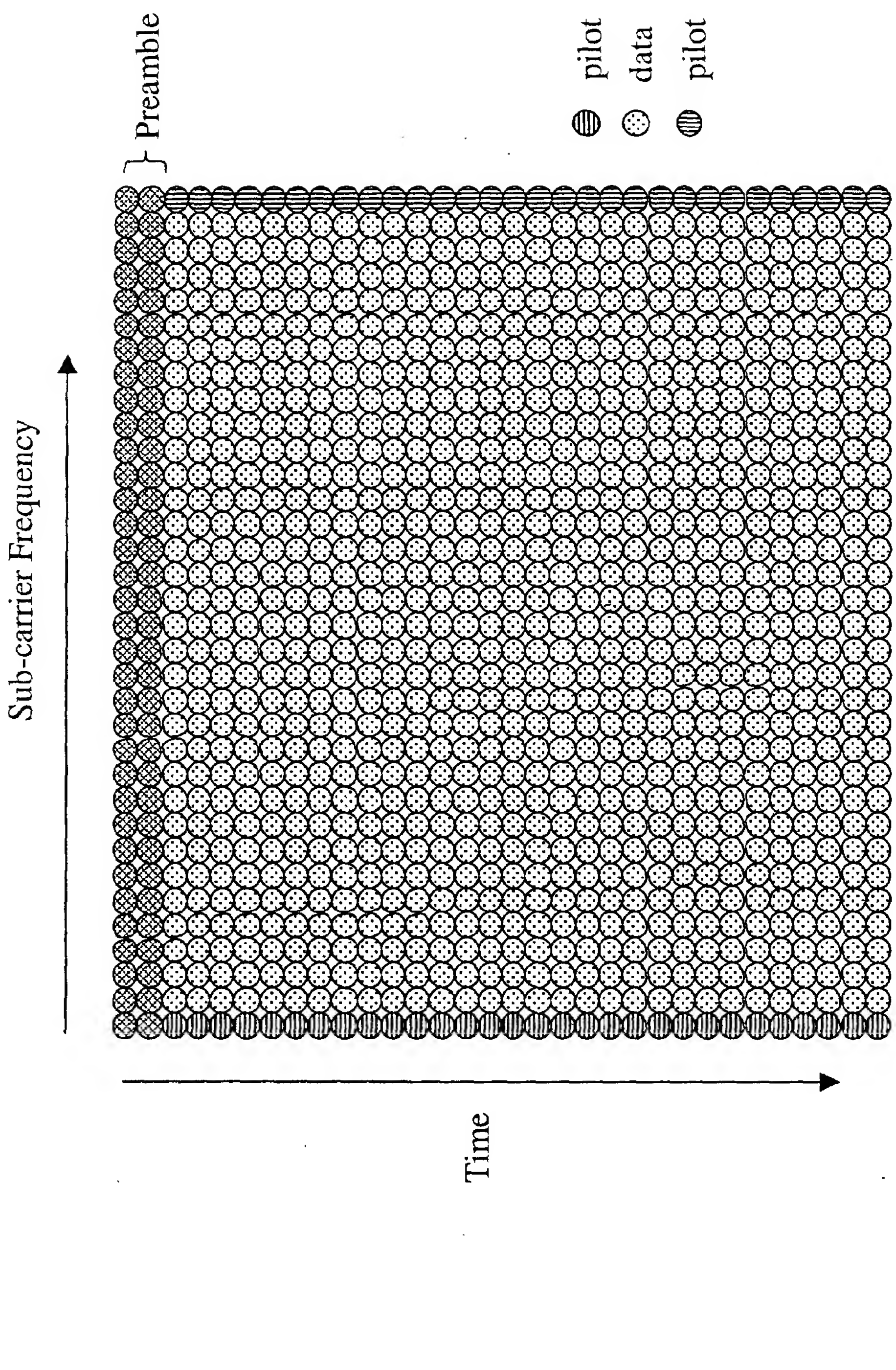

FIG. 3

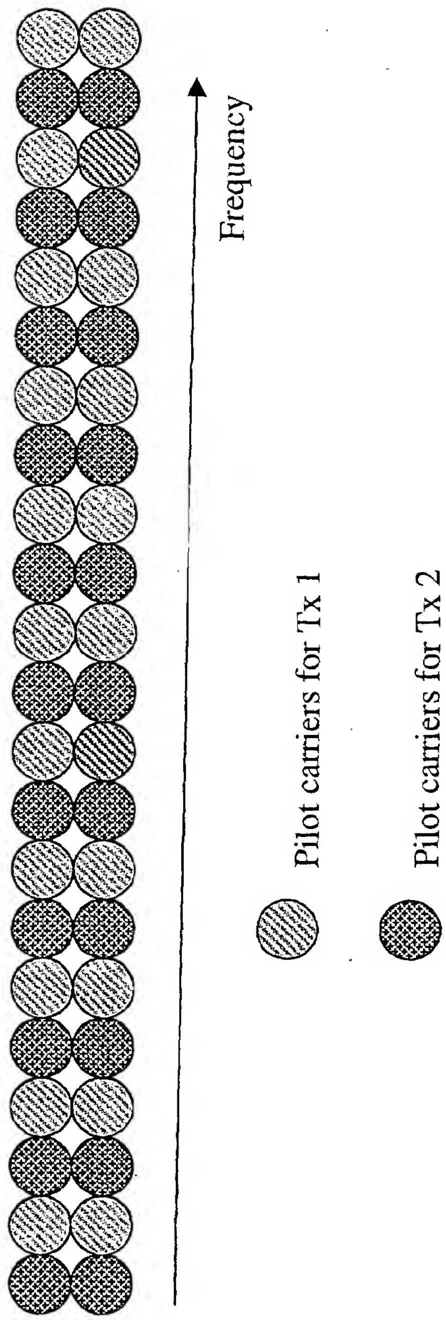

FIG. 4

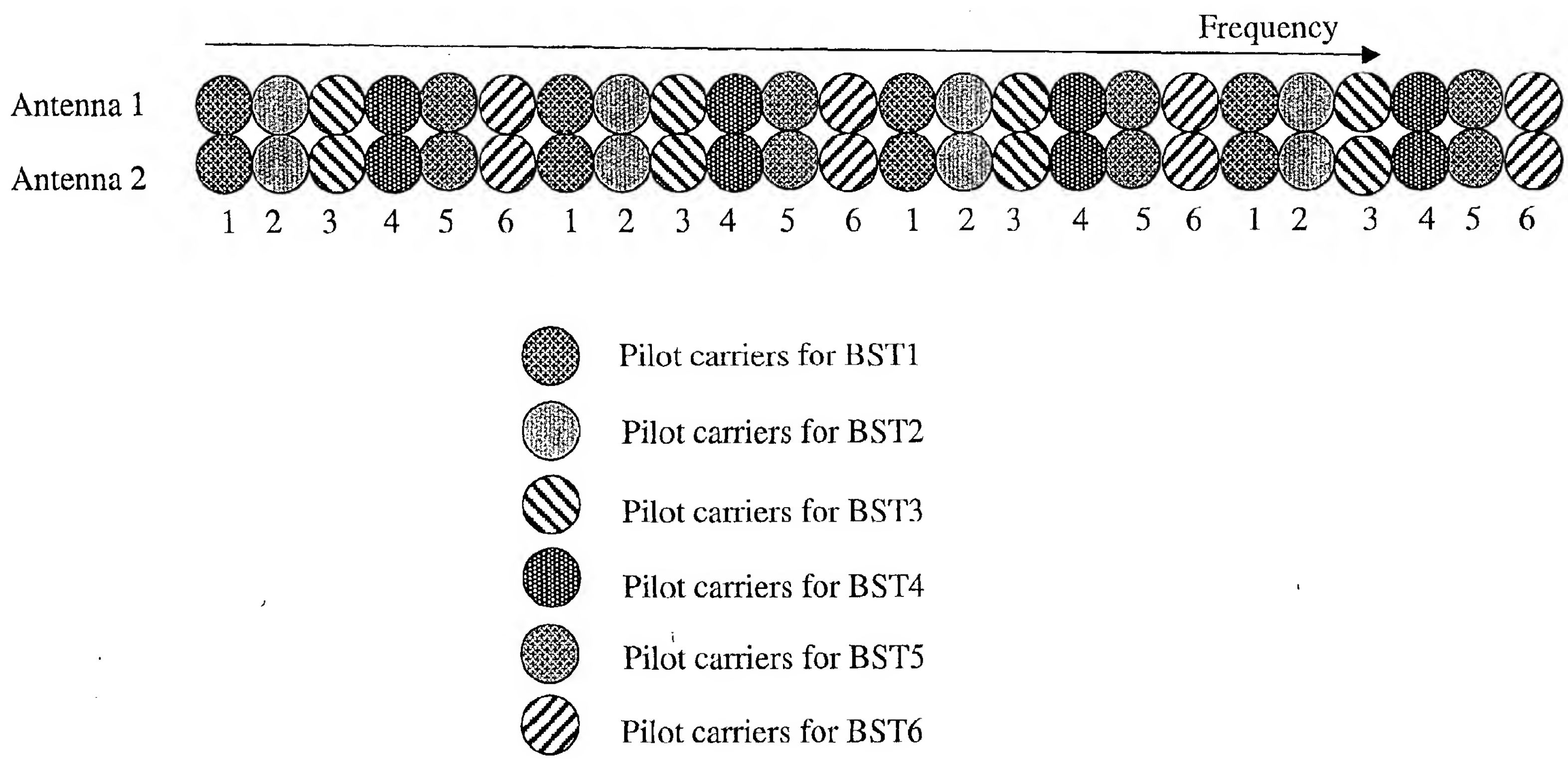
Frequency
Antenna 1
Antenna 2
1 2 3 4 5 6 1 2 3 4 5 6 1 2 3 4 5 6 1 2 3 4 5 6
Pilot carriers for BST1
Pilot carriers for BST2
Pilot carriers for BST3
Pilot carriers for BST4
Pilot carriers for BST5
Pilot carriers for BST6

FIG. 5

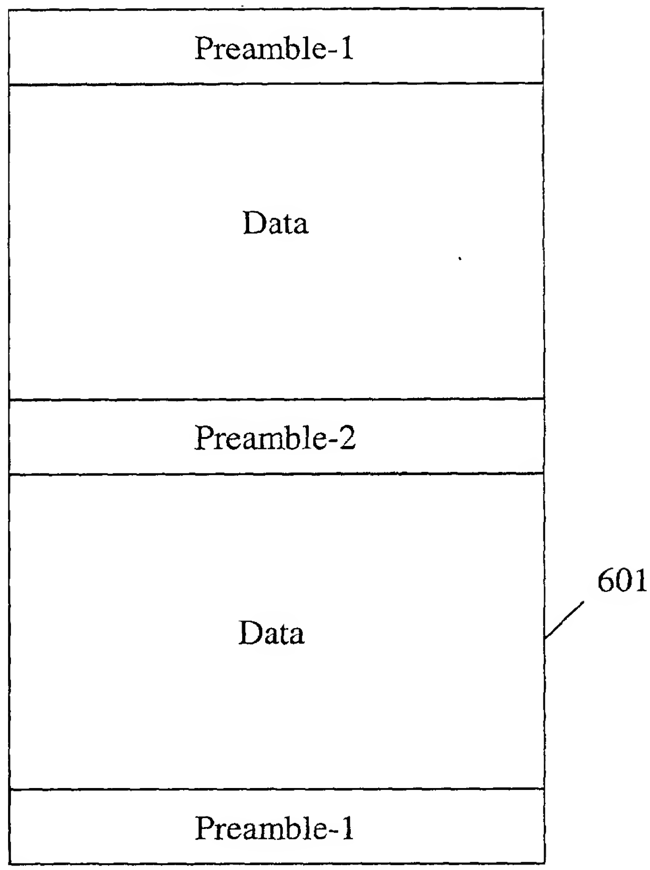

FIG. 6

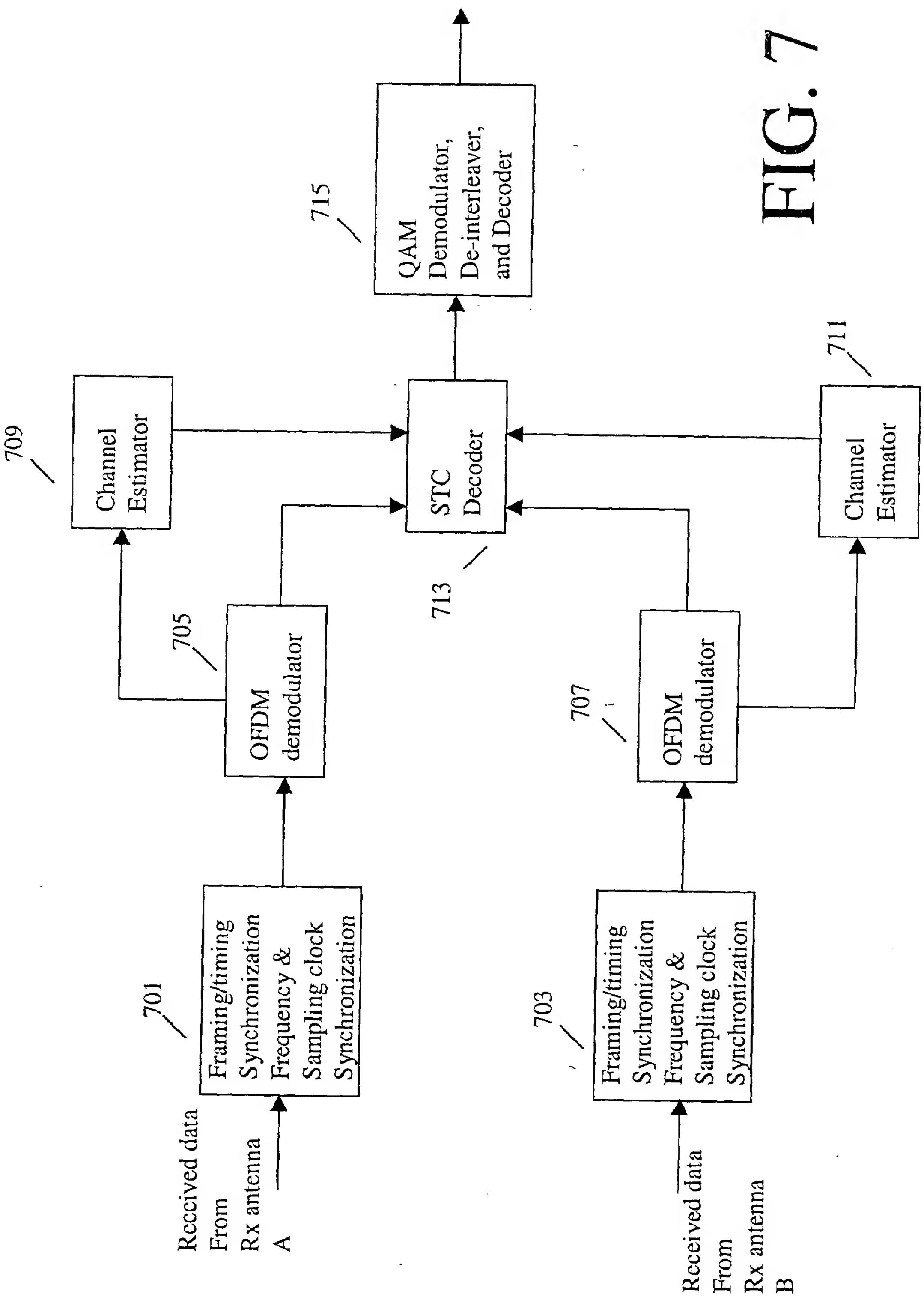
Received data From Rx antenna A
701
Framing/timing Synchronization Frequency & Sampling clock Synchronization
705
OFDM demodulator
709
Channel Estimator
Received data From Rx antenna B
703
Framing/timing Synchronization Frequency & Sampling clock Synchronization
707
OFDM demodulator
711
Channel Estimator
713
STC Decoder
715
QAM Demodulator, De-interleaver, and Decoder

FIG. 7

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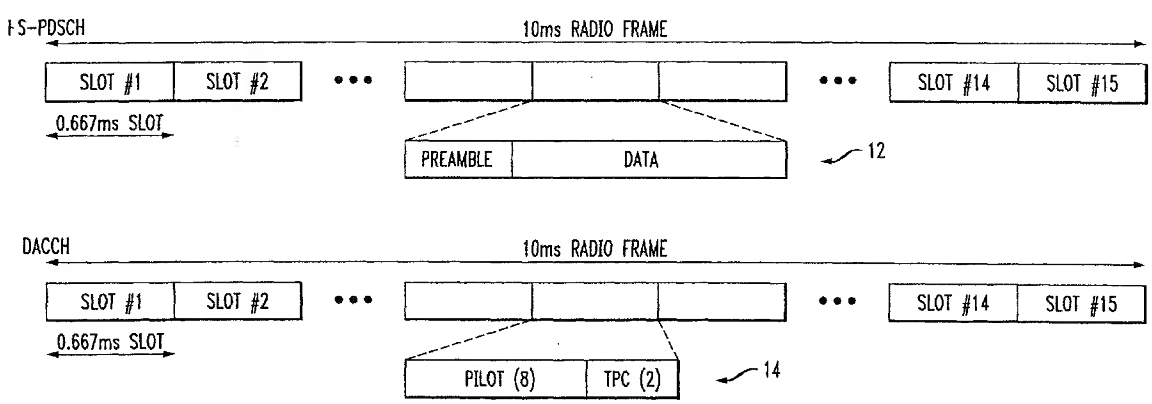
(54) Downlink and uplink channel structures for downlink shared channel system

(57) An uplink and downlink channel structure supports a shared downlink data channel. The new structure accommodates advanced physical and Medium Access Control (MAC) layer techniques, such as incremental redundancy (IR), fast adaptation to channel conditions, and multiple input multiple output (MIMO) antenna configuration. The proposed changes are intended to lead to a downlink structure that achieves higher spectral efficiency for the packet oriented services over then shared downlink channel. Additionally, the new structure uses the base station transmit power information and of the channelization (OVSF) code space more efficiently.

FIG. 2



EP 1 207 635 A1

Description

BACKGROUND OF THE INVENTION

1. Field of The Invention

[0001] This invention relates to wireless communications and, more particularly, to downlink and uplink channel structures to support a downlink shared channel system.

2. Description of Related Art

[0002] Wireless communications systems include conventional cellular communication systems which comprise a number of cell sites or base stations, geographically distributed to support transmission and receipt of communication signals to and from wireless units which may actually be stationary or fixed. Each cell site handles voice communications over a particular region called a cell, and the overall coverage area for the cellular communication system is defined by the union of cells for all of the cell sites, where the coverage areas for nearby cell sites overlap to some degree to ensure (if possible) contiguous communications coverage within the outer boundaries of the system's coverage area.

[0003] When active, a wireless unit receives signals from at least one base station or cell site over a forward link or downlink and transmits signals to (at least) one cell site or base station over a reverse link or uplink. There are many different schemes for defining wireless links or channels for a cellular communication system, including TDMA (time-division multiple access), FDMA (frequency-division multiple access), and CDMA (code-division multiple access) schemes. In CDMA communications, different wireless channels are distinguished by different codes or sequences that are used to encode different information streams, which may then be modulated at one or more different carrier frequencies for simultaneous transmission. A receiver can recover a particular information stream from a received signal using the appropriate code or sequence to decode the received signal.

[0004] Due to the delay-intolerant nature of voice communication, wireless units in conventional cellular systems transmit and receive over dedicated links between a wireless unit and a base station. Generally, each active wireless unit requires the assignment of a dedicated link on the forward link and a dedicated link on the reverse link. Traditional data applications are typically bursty and, unlike voice communications, relatively delay tolerant. As such, using dedicated links to transmit data is an inefficient use of network resources. Wireless communications systems are evolving that will support a variety of data services, such as wireless web browsing.

[0005] In the Universal Mobile Telecommunications System (UMTS), wireless units communicate with a base station over dedicated channels. To provide efficient wireless data communications, UMTS uses a downlink shared channel which can be shared by a plurality of wireless units to receive data. To improve system throughput, the system provides the wireless unit with the best reported rate access to the shared channel.

SUMMARY OF THE INVENTION

[0006] The present invention is an uplink and downlink structure to support a shared downlink data channel. The new structure accommodates physical and Medium Access Control (MAC) layer techniques, such as incremental redundancy (IR), fast adaptation to channel conditions, and multiple input multiple output (MIMO) antenna configuration. The proposed changes are intended to lead to a downlink structure that achieves higher spectral efficiency for the packet oriented services over the shared downlink channel. Additionally, the new structure uses the base station transmit power information and the channelization (OVSF) code space more efficiently.

BRIEF DESCRIPTION OF THE DRAWINGS

[0007] Other aspects and advantages of the present invention may become apparent upon reading the following detailed description and upon reference to the drawings in which:

FIG. 1 shows a general diagram of an embodiment of the uplink dedicated physical control channel (UL DPCCH) frame structure according to principles of the present invention; and
FIG. 2 shows a general diagram of an embodiment of the downlink shared channel (HS-DSCH) and the downlink associated control channel (DACCH) acccording to principles of the present invention.

DETAILED DESCRIPTION

[0008] Illustrative embodiments of the downlink and uplink channel structures are described with respect to a UMTS

system where the downlink shared channel is formed using at least one and more likely a plurality of channelization codes. The downlink shared channel is time-division multiplexed, being divided into 10 millisecond frames of 15 slots of .667 milliseconds. Wireless units provide rate and antenna feedback from which the base station decides whether to provide the wireless unit to the shared data channel.

[0009] Key aspects supported by the uplink and downlink channel structures are outlined here and a more detailed embodiment is provided below.

1. Explicit Rate and Antenna Information from the wireless unit:
Explicit Rate feedback from the wireless unit is preferred over sending Signal to Interference Ratio (SIR) measurements to the base station for the following reasons:

- The wireless unit is aware of the circuit switched or dedicated channel loading (by reading the available power fraction on the BEACCH described below) on all active code channels and uses this information in deciding the rate as well as the best serving cell during fast cell selection described below.
- The wireless unit is aware of the power fraction (the fraction of the power that is currently occupied by all other downlink dedicated physical channels) as well as current and future (RAI round-trip delay) data activity of all active connections or code channels with which the wireless unit is in soft handoff (simultaneously receiving from different base stations) as well as the path losses from each of these active soft handoff connections and uses this information in estimating the total expected interference better. This permits a more accurate determination of the rate that is important for the performance and fair scheduling of cell edge wireless units that are dominated by interferers.

There are multiple antenna configurations supported by embodiments of the channel structure. In certain embodiments, the RAI field in the UL DPCCH contains rate as well as information which can select the appropriate antenna or antenna configuration. Depending on the rate information (RI), the transmission/reception can be achieved via either multiple input/single output or multiple input/multiple antenna configurations.

2. Preamble Based User Identification on the downlink shared channel:
The downlink shared channel is associated with a Downlink Dedicated Channel (DCH). This dedicated channel is used to identify the user that is scheduled for transmission in the shared channel. A preamble or a local or temporary identifier is a more efficient scheme to identify a user compared to a unique user identification. Larger spreading factor (SF) channelization codes can then be used, hence saving the downlink channelization code space.

3. Asymmetric Downlink and Uplink Transmission Time Interval
From a frame-fill efficiency perspective, it is preferable to have one slot (0.667ms) as the granularity for the downlink shared channel frame. This is because at data rates in excess of 10 Mbps and with typical Internet packet sizes, frame sizes greater than one slot could lead to considerable inefficiency. However, the link quality (or rate information) feedback from the wireless unit does not have to be at the rate of once per slot. The typical rate at which channel quality changes is slower than that of one slot. Besides, the overhead associated with per slot feedback of the link quality (or rate) would be excessive. Therefore, in certain embodiments, the uplink transmission time interval (TTI) is based on 3 slots or 2 ms.

4. Fast Cell-site Selection (FCS):
To support fast cell site selection, the wireless unit selects the best cell for its downlink transmission every TTI. Selecting the best cell based purely on signal strength measurement at the wireless unit unaware of the loading in the surrounding cells can result in selecting a cell that is heavily loaded. In later section a Downlink Broadcast channel is defined, among other functions, to support the wireless unit fast cell site selection. In this Downlink Broadcast channel, both the base station transmitter power fraction and an indication of the orthogonal variable spreading factor (OVSF) or channelization code space are sent. In addition, an OVSF code cover based approach for signalling the preferred cell site is proposed. The advantage with an OVSF code cover based approach over a coded bit field based approach is that it is much faster to detect. Thus, fields defined in the uplink dedicated physical control channel for cell ID can be avoided.

[0010] FIG. 1 shows an embodiment of the uplink (UL) dedicated physical control channel (DPCCH) frame structure 10 that supports a downlink shared channel. As discussed above, from a frame-fill efficiency perspective, it is best to retain a single-slot (0.667ms) granularity for the downlink shared channel physical frame. However, for the uplink, the constraint may be relaxed. Here, a goal is to keep the feedback rate (bps) required to support the downlink shared channel adequately low. The higher the feedback rate, the greater the noise rise and consequently, the greater the reduction in uplink capacity for dedicated channels. Of a 10 ms radio frame divided into 15 slots of .67 ms each, a three-slot (2ms) granularity for feedback of measured downlink quality information achieves a good trade-off between

link quality tracking and feedback overhead.

[0011] The DPCCH spreading factor is lowered from 256 to 128. This allows for 20 coded bits per slot, and downlink shared channel-related control information is readily accommodated via the uplink dedicated physical control channel (UL DPCCH)-related control information. As before, each slot has 2560 chips. Two new fields are defined: Rate and Antenna Information field (RAI) and an acknowledgement (ACK)/negative acknowledgement (NACK) field. Other fields that already exist in current known versions of uplink DPCCH are retained: Pilot, TFCI, FBI and TPC. Pilot bits (5 per slot) will be used for coherent demodulation, TFCI bits (2 per slot) indicate the transport frame configuration or format of the associated uplink (UL) dedicated physical data channel (DPDCH) (if any), the FBI bits (2 per slot) indicate antenna weights and/or the site chosen for the downlink dedicated physical data channel (DPDCH) (if any), not the DSCH, and the TPC bits (1 per slot) are used for downlink power control on all the dedicated downlink channels associated with the user (again, not the DSCH).

[0012] Explicit signalling of Rate and Antenna Information (RAI) from the wireless unit is preferred as compared to signalling the SIR estimate for reasons cited above in Paragraph 1 of the key aspects. Towards this end a 5-bit Rate and Antenna Information Field is defined. This will comprise of a 4 bit Rate Information (RI) part that allows the wireless unit to select from one of sixteen possible AMC (Adaptive Modulation and Coding) states and a 1-bit antenna indication (AI) field. The role of the AI field can be made dependent on the RI bits. For example, the allowed rates could be partitioned into two disjoint sets, high and low. If the RI field indicates a rate from the high set then the AI field could signal MIMO or non-MIMO reception, whereas if the RI field indicates a rate from the low set, then the AI field could be purely antenna selection. The wireless unit determines the RAI field based on downlink quality estimates, available downlink shared channel power, available downlink orthogonal variable spreading factor (OVSF) code space and predicted neighbor cell loading. A rate 1/3 block code could then be used to map the 5 RAI bits to 15 coded symbols. These 15 coded symbols are carried over three time slots. Such a structure supports the concept of a rate feedback rate of every three slots. Depending on the embodiment, the feedback rate can be slower than the rate of adapting the downlink shared channel in accordance with the rate information.

[0013] A single bit ACK/NACK field is defined in support of Incremental Redundancy (IR) and it indicates whether the received packet was in error or not. For example, if X bits are to be transmitted, redundancy is coded into the X bits to produce 3*X bits to be sent. In incremental redundancy, the first X bits are sent, and if a NACK is received, the next X bits are sent. If a NACK is again received, then the final X bits are sent which should be able to be decoded. The rate of ACK/NACK signalling is once per slot i.e. once every 0.667 ms as opposed to the RAI field which is defined over three slots i.e. 3*0.667=2ms. The ACK bit is repeated five times to form five coded symbols and transmitted over 0.667 ms duration. When the wireless unit does not have a transmission to acknowledge, the ACK/NACK field is ignored by the base station or could be gated OFF. Transmission of 5 RAI bits every 3 slots and 1 ACK/NACK bit per slot results in a total uplink (UL) feedback rate for High Speed Downlink Packet Access (HSDPA) of about 4kbps. This compares well with an average bit rate of around 6kbps for 12.2Kbps Adaptive Multirate (AMR) for voice service with an activity factor of 0.5.

[0014] Fast cell site selection will be based on using OVSF code covers for the RAI field. Instead of using the uplink DPCCH channelization code, the cell is identified by a different OVSF code from a limited pre-determined set of OVSF codes of spreading factor 128.

[0015] FIG. 2 shows the frame structure of an embodiment of the downlink shared channel (HS-DSCH)12 and the downlink associated control channel (DACCH) 14. It is known that a shorter Transmission Time Interval (TTI) provides certain advantages. Smaller granularity for the downlink shared channel TTI provides the following advantages:

- Better base station adaptation of the downlink shared channel due to the formation of smaller size packets. This is essential for the higher rates of the downlink shared channel and leads to higher frame-fill efficiency. Note, internet packet sizes vary from 60 bytes-1500 bytes.
- Better adaptation to the channel conditions when combined with efficient and fast scheduling. By using smaller TTI or packet sizes, changes in the channel can be fed back and applied faster on the downlink.

Therefore, it is proposed that the downlink shared channel 12 meet the the following two basic definitions:

- The downlink shared channel TTI equals 1 slot interval, TTI =1 x Ts. The slot duration is Ts = 0.667ms.
- This new TTI is the minimum time interval for which the downlink shared channel resource is allocated to a wireless unit. The actual time for transmission depends on the Incremental Redundancy (IR) performance.

[0016] Known systems have a downlink shared channel associated with a Downlink Dedicated Channel (DCH) which maps into a physical dedicated channel (PDCH). The DCH is used to point to the downlink shared channel. The DCH indicates to the wireless unit when it should decode the downlink shared channel (DSCH) and the associated spreading code information. This method does not make the most efficient use of the available channelization code space.

[0017] The new DACCH carries the TPC bits that implement uplink power control. In addition, the Pilot bits are time-multiplexed in this channel as well. For each wireless unit, their DACCH are code multiplexed with an OVSF code of spreading factor (SF) 512, resulting in considerable code space saving in the downlink. Furthermore, since this is a dedicated channel, it is power controlled from the TPC bits sent in the UL DPCCH.

[0018] The definition of this new channel replaces the dedicated pointer channel (DPTRCH) known in the art. Similarly, new functionality is defined for the DSCH in the next section. The DACCH fields are shown in Table 1.

Table 1.

DACCH Fields				
Channel	SF	Bits/Slot	Pilot (Bits/Slot)	TPC (Bits/Slot)
DACCH	512	10	8	2

[0019] The HS-DSCH uses multi-code transmission using the available channelization code space. In this channel, traffic data and a preamble, per TTI, are time multiplexed within the downlink shared channel frame. As described above, a shorter TTI that is equal to 1 slot is proposed. The preamble field duration per TTI is not fixed and is determined by RAI field in the UL DPCCH. The preamble contains fields that declare:

- MAC user ID to which the TTI belongs.
- Asynchronous Incremental Redundancy (AIR)

The use of preamble within the HS-DSCH TTIs alleviates the use of additional code-multiplexed channels that will have to carry the various control fields. The preamble solution preserves the OVSF code space, and reduces decoding latencies. The preamble is of variable length depending on the UL DPCCH RI field decoding. The wireless unit is then aware of the preamble length.

Table 2:

HS-DSCH Fields	
Preamble (bits)	**Data (bits)**
Variable length dependent on the UL DPCCH RI field decoding.	Variable length dependent on the decoding of the UL DPCCH RI field decoding.

[0020] A new Beacon Control Channel (BEACCH) is defined. The coexistence of the downlink shared channel (HS-DSCH) with circuit-switched or dedicated downlink channels requires that the downlink power available to the HS-DSCH users, as well as the subset of the channelization code space available for multicode transmission be broadcasted to all HS-DSCH users. This information is updated at the following rates and are also shown in Table 3:

- The power fraction (PF) available for the HS-DSCH is updated every TTI (1 slot). This update rate is required in order to follow the power control rate of the voice channels.
- The downlink Activity Indicator (DAI) available for HS-DSCH is updated every TTI (1 slot). This field indicates the future (RAI round-trip delay) data activity on the HS-DSCH.
- The available OVSF Code Space (OCS) for the HS-DSCH once every 10ms frame, the rate by which voice users enter or exit the physical layer.

Table 3.

BEACCH Channel Field Structure				
Channel	**SF**	**Bits/Slot**	**Power Fraction & DAI (Bits/Slot)**	**OCS (Bits/Slot)**
BEACCH	256	20	14	6

[0021] In addition to the embodiment(s) described above, the uplink and downlink channel structure has been described for use in a UMTS system where the downlink data channel is a shared, time division multiplexed channel made up of at least one channelization code. The uplink and/or downlink channel structure according to the principles of the present invention can be used with different cellular systems and uplink and/or downlink configurations which omit and/or add components and/or use variations or portions of the described system. For example, the rate informa-

tion fed back to the base station can include a code sequence which maps to a particular configuration, including coding, modulation and/or antennas, for the base station to adapt the downlink data channel for communication with the wireless unit. Alternatively, the rate information could include a rate or other information the base station can use to calculate a rate or appropriate configuration to communicate over the data channel.

[0022] It should be understood that the system and portions thereof and of the described uplink and/or downlink channel structure can be implemented in different locations, such as the wireless unit, the base station, a base station controller and/or mobile switching center. Moreover, logic or hardware required to implement and use the uplink and/ or downlink channel structure can be implemented in application specific integrated circuits, software-driven processing circuitry, firmware, programmable logic devices, hardware, discrete components or arrangements of the above components as would be understood by one of ordinary skill in the art with the benefit of this disclosure. What has been described is merely illustrative of the application of the principles of the present invention. Those skilled in the art will readily recognize that these and various other modifications, arrangements and methods can be made to the present invention without strictly following the exemplary applications illustrated and described herein and without departing from the spirit and scope of the present invention.

Claims

1. A method comprising the steps of:

transmitting at a rate on a downlink shared channel having slots with each slot including a preamble locally identifying a wireless unit and data;
transmitting on a dedicated control channel associated with a dedicated downlink channel including power control bits and pilot bits; and
transmitting on a beacon channel a power fraction information, downlink activity information and code space information.

2. The method of claim 1 comprising:

receiving on an uplink dedicated physical control channel, rate information for transmitting on said downlink shared channel, and an acknowledgment information indicating the receipt of data.

3. A method comprising the steps of:

receiving at a rate on a downlink shared channel having slots with each slot including a preamble locally identifying a wireless unit and data;
receiving on a dedicated control channel associated with a dedicated downlink channel including power control bits and pilot bits; and
receiving on a beacon channel a power fraction information, downlink activity information and code space information.

4. The method of claim 1 comprising:

transmitting a rate on an uplink dedicated physical control channal, rate information calculated using said power fraction information, said downlink activity information and said code space and acknowledgment information indicating the receipt of data.

FIG. 1

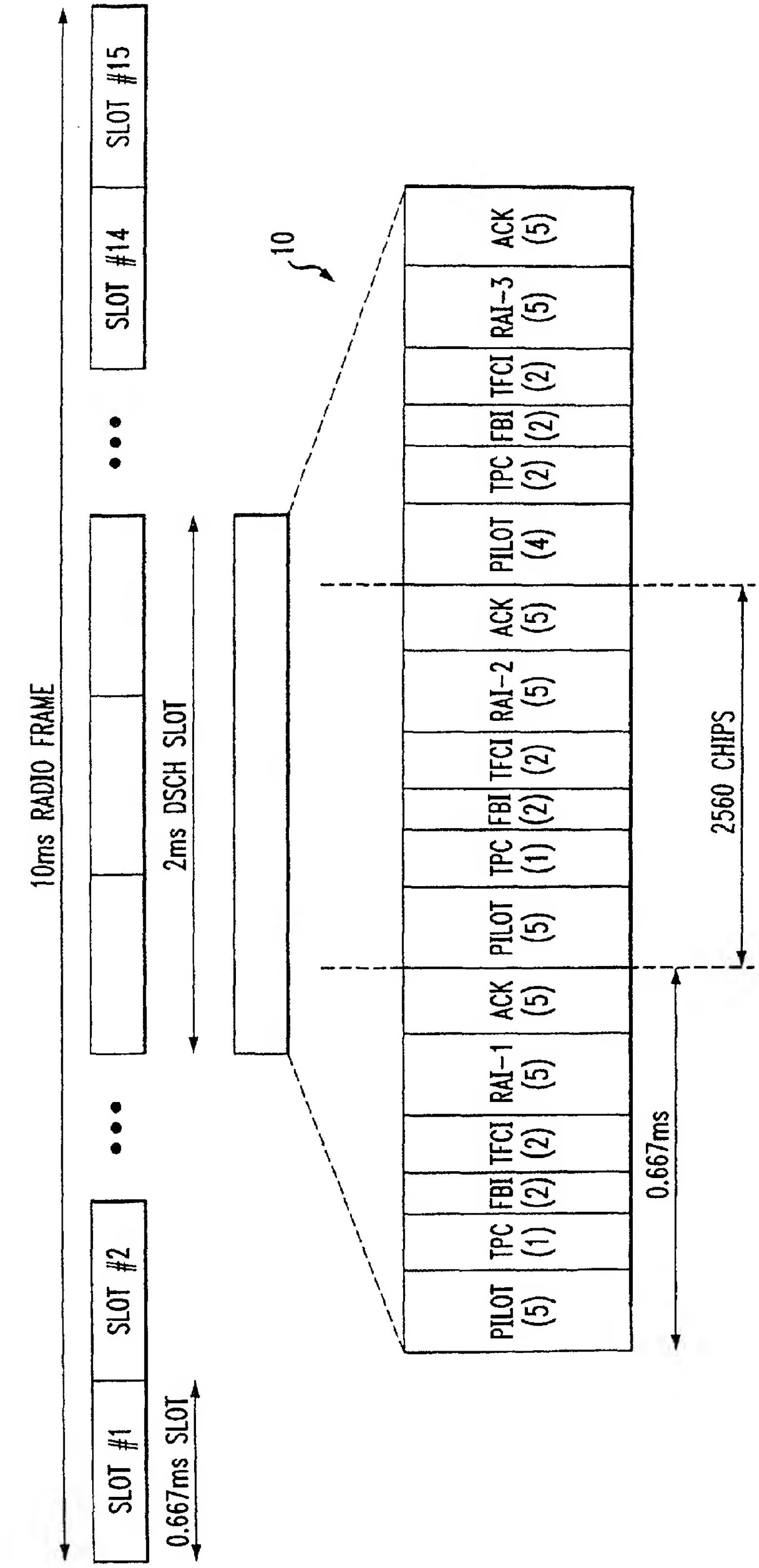
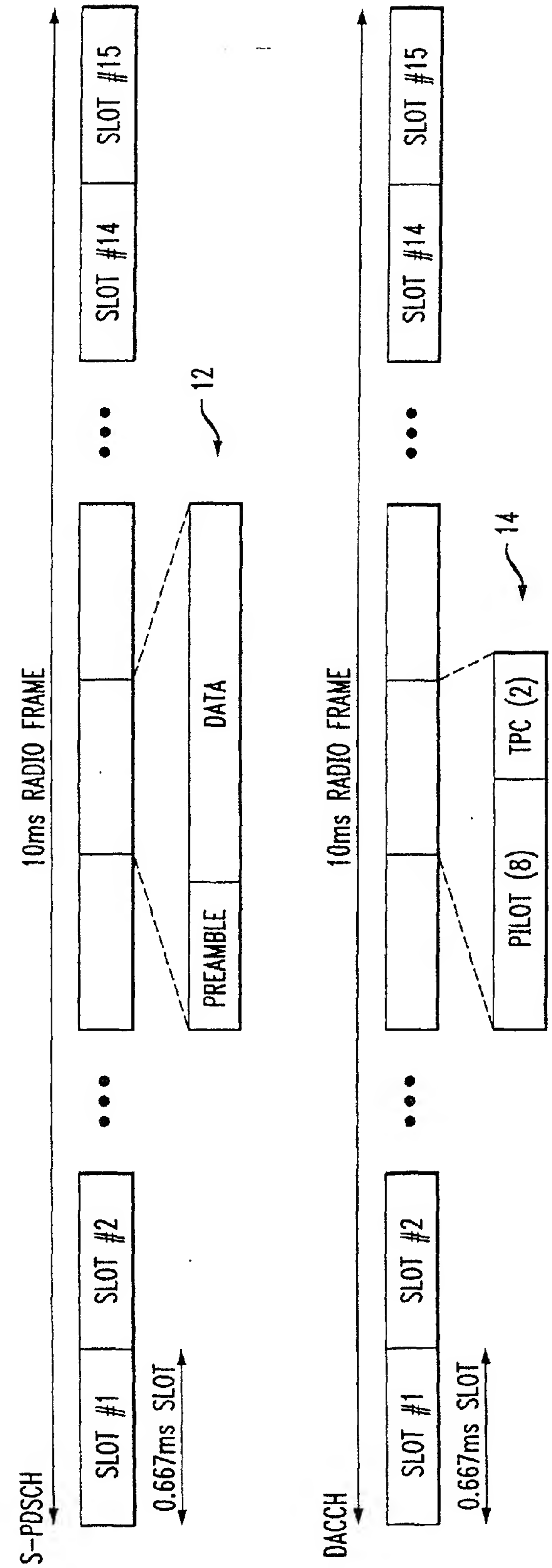

10ms RADIO FRAME
SLOT #1
SLOT #2
SLOT #14
SLOT #15
0.667ms SLOT
2ms DSCH SLOT
10
PILOT (5)
TPC (1)
FBI (2)
TFCI (2)
RAI-1 (5)
ACK (5)
PILOT (5)
TPC (1)
FBI (2)
TFCI (2)
RAI-2 (5)
ACK (5)
PILOT (4)
TPC (2)
FBI (2)
TFCI (2)
RAI-3 (5)
ACK (5)
0.667ms
2560 CHIPS

FIG. 2


HS-PDSCH
10ms RADIO FRAME
SLOT #1
SLOT #2
SLOT #14
SLOT #15
0.667ms SLOT
PREAMBLE
DATA
12
DACCH
10ms RADIO FRAME
SLOT #1
SLOT #2
SLOT #14
SLOT #15
0.667ms SLOT
PILOT (8)
TPC (2)
14

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 30 4724

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
P,X	LUCENT TECHNOLOGIES: "TSGR1#17(00)1384" TSG-RAN WORKING GROUP 1, 'Online! 21 - 24 November 2000, XP002175651 Stockholm, Sweden Retrieved from the Internet: <URL:http://www.3gpp.org/ftp/tsg_ran/WG1_RL1/TSGR1_17/Docs/PDFs/R1-00-1384.pdf> 'retrieved on 2001-08-23! * The whole document *	1-4	H04B7/26 H04B7/005
A	EP 0 954 118 A (ROKE MANOR RESEARCH) 3 November 1999 (1999-11-03) * column 3, line 37 - column 4, line 52; claims 1-5 *	1-4	
A	EP 0 993 128 A (MOTOROLA INC) 12 April 2000 (2000-04-12) * column 3, line 5 - line 35 *	1-4	
A	US 6 009 091 A (STEWART KENNETH A ET AL) 28 December 1999 (1999-12-28) * column 1, line 41 - line 63 *	1-4	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04B H04Q

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	3 September 2001	Sorrentino, A

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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A : technological background
O : non-written disclosure
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T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 01 30 4724

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report.
The members are as contained in the European Patent Office EDP file on
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03-09-2001

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EP 0954118	A	03-11-1999	NONE		
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EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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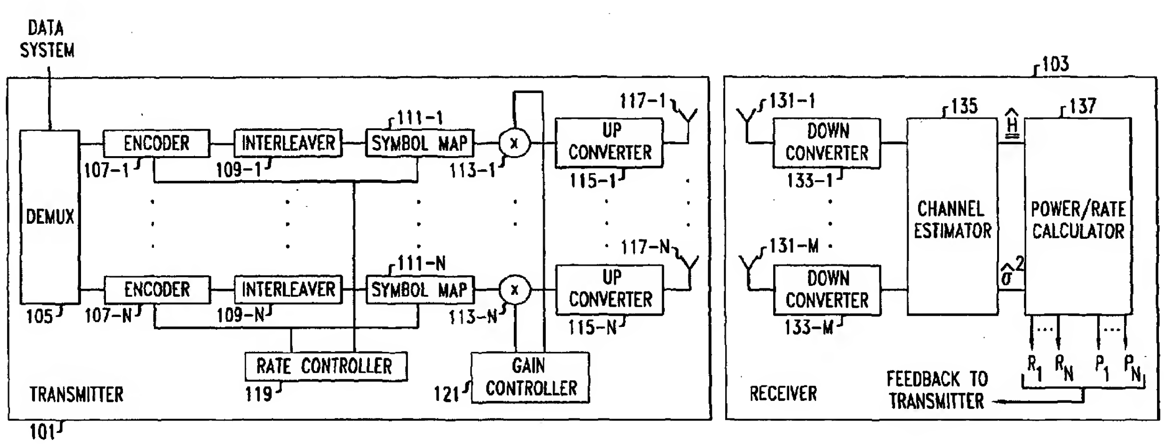
(54) Feedback technique for wireless systems with multiple transmit and receive antennas

(57) In a wireless communication system using multiple antennas at the transmitter and multiple antennas at the receiver, a so called multiple-input multiple-output (MIMO) system, a substantial improvement in capacity over the case of no feedback can be achieved using considerably less bandwidth than is required to feedback the channel estimate or channel statistics, by supplying as feedback for each data substream of an overall data stream an indicator of a rate and/or an indicator of a gain for transmission of that data substream. The indicator of the rate and/or the indicator of the gain may be the rate and/or gain directly or an encoded representation of the rate and/or gain. Typically, the best performance is achieved if indicators of both the rate and the gain are fed back. If the invention is implemented for wireless systems, then typically for each data substream there is a separate, independent antenna. The rate and the gain are computed as a function of a channel estimate which is developed at the receiver.. Advantageously, the transmitter may employ only one-dimensional data coding for each of the data substreams.

FIG. 1



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Description

Technical Field

[0001] This invention relates to the art of wireless communications, and more particularly, to wireless communication systems using multiple antennas at the transmitter and multiple antennas at the receiver, so called multiple-input multiple-output (MIMO) systems.

Background of the Invention

[0002] It is known in the art that multiple-input multiple-output (MIMO) systems can achieve dramatically improved capacity as compared to single antenna, i.e., single antenna to single antenna or multiple antenna to single antenna, systems. It is also known in the art that if a channel estimate or channel statistics based on the channel estimate are fed back to the transmitter, then the throughput of the channel can be improved with respect to an identically configured system but without feedback. However, because in MIMO systems the overall channel is actually made up of multiple channels, with one channel for each transmit and receive pairing, such feedback requires considerable bandwidth, and it is undesirable to dedicate so much bandwidth to feedback.

Summary of the Invention

[0003] In a MIMO system, a substantial improvement over the case of no feedback can be achieved using considerably less bandwidth than is required to feedback the channel estimate or channel statistics, in accordance with the principles of the invention, by supplying as feedback for each data substream of an overall data stream an indicator of a rate and/or an indicator of a gain for transmission of that data substream. The indicator of the rate and/or the indicator of the gain may be the rate and/or gain directly or an encoded representation of the rate and/or gain. Typically, the best performance is achieved if indicators of both the rate and the gain are fed back. If the invention is implemented for wireless systems, then typically for each data substream there is a separate, independent antenna. The rate and the gain are computed as a function of a channel estimate which is developed at the receiver. Advantageously, the transmitter may employ only one-dimensional data coding for each of the data substreams.

Brief Description of the Drawing

[0004] In the drawing:

FIG. 1 shows an exemplary multiple-input multiple-output (MIMO) system arranged in accordance with the principles of the invention so as to achieve dramatically improved capacity as compared to single antenna systems; and

FIG. 2 shows an exemplary process for determining rates and powers for a system with N transmit substreams and M receive branches, in accordance with an aspect of the invention.

Detailed Description

[0005] The following merely illustrates the principles of the invention. It will thus be appreciated that those skilled in the art will be able to devise various arrangements which, although not explicitly described or shown herein, embody the principles of the invention and are included within its spirit and scope. Furthermore, all examples and conditional language recited herein are principally intended expressly to be only for pedagogical purposes to aid the reader in understanding the principles of the invention and the concepts contributed by the inventor(s) to furthering the art, and are to be construed as being without limitation to such specifically recited examples and conditions. Moreover, all statements herein reciting principles, aspects, and embodiments of the invention, as well as specific examples thereof, are intended to encompass both structural and functional equivalents thereof. Additionally, it is intended that such equivalents include both currently known equivalents as well as equivalents developed in the future, i.e., any elements developed that perform the same function, regardless of structure.

[0006] Thus, for example, it will be appreciated by those skilled in the art that the block diagrams herein represent conceptual views of illustrative circuitry embodying the principles of the invention. Similarly, it will be appreciated that any flow charts, flow diagrams, state transition diagrams, pseudocode, and the like represent various processes which may be substantially represented in computer readable medium and so executed by a computer or processor, whether or not such computer or processor is explicitly shown.

[0007] The functions of the various elements shown in the FIGs., including functional blocks labeled as "processors" may be provided through the use of dedicated hardware as well as hardware capable of executing software in association with appropriate software. When provided by a processor, the functions may be provided by a single dedicated processor, by a single shared processor, or by a plurality of individual processors, some of which may be shared. Moreover, explicit use of the term "processor" or "controller" should not be construed to refer exclusively to hardware capable of executing software. and may implicitly include, without limitation, digital signal processor (DSP) hardware, read-only memory (ROM) for storing software, random access memory (RAM), and non-volatile storage. Other hardware, conventional and/or custom, may also be included. Similarly, any switches shown in the FIGS. are conceptual only. Their function may be carried out through the operation of program logic, through dedicat-

ed logic, through the interaction of program control and dedicated logic, or even manually, the particular technique being selectable by the implementor as more specifically understood from the context.

[0008] In the claims hereof any element expressed as a means for performing a specified function is intended to encompass any way of performing that function including, for example, a) a combination of circuit elements which performs that function or b) software in any form, including, therefore, firmware, microcode or the like, combined with appropriate circuitry for executing that software to perform the function. The invention as defined by such claims resides in the fact that the functionalities provided by the various recited means are combined and brought together in the manner which the claims call for. Applicant thus regards any means which can provide those functionalities as equivalent as those shown herein.

[0009] FIG. 1 shows an exemplary multiple-input multiple-output (MIMO) system arranged in accordance with the principles of the invention so as to achieve dramatically improved capacity as compared to single antenna systems. In particular, FIG. 1 shows transmitter 101 and receiver 103. Transmitter 101 includes a) demultiplexer (demux) 105; b) encoders 107, including encoders 107-1 through 107-N; c) interleavers 109, including interleavers 109-1 through 109-N; d) symbol mappers 111, including symbol mappers 111-1 through 111-N; e) gain multipliers 113, including gain multipliers 113-1 through 113-N; f) optional upconverters 115, including optional upconverters 115-1 through 115-N; g) optional transmit antennas 117, including optional transmit antennas 117-1 through 117-N; h) rate controller 119; and i) gain controller 121. Receiver 103 includes a) optional receive antennas 131, including optional receive antennas 131-1 through 131-M; b) optional downconverters 133, including optional downconverters 133-1 through 133-M; c) channel estimator 135; and d) power/rate calculator 137.

[0010] Demultiplexer 105 receives as an input an overall data stream, which is the data to be transmitted, and divides it into N data substreams, each to be processed along an independent transmit path and then transmitted.

[0011] Each of encoders 107 applies channel coding to the respective data substream it receives so as to increase the redundancy of the data substream. This facilitates error recovery at the receiver should errors occur. In accordance with an aspect of the invention, the type of channel coding used is a function of the rate, or an indicator thereof, that is fed back from the receiver. This function may be implemented using a lookup table given an indicator of the rate that is fed back, and is typically implemented by rate controller 119. Those of ordinary skill in the art will readily appreciate how to arrange such a function given the particular details of the system being implemented, e.g., the channel statistics, the number of substreams employed, and the like. The type of channel coding employed determines the particular amount of redundancy in the encoded data substream, and it is noted that the amount of redundancy is known as the code rate. Each of encoders 107 may use a channel coding that is independent of the channel coding used by any other of encoders 107, and each may receive an independently specified rate.

[0012] Interleavers 109 are conventional in nature and each rearranges the bits of the encoded data substream it receives to provide protection against channel fades.

[0013] Each of symbol mappers 111 maps the bits of the interleaved encoded channel substream that it receives to a point in a constellation. In accordance with an aspect of the invention, the particular constellation employed is selected as a function of the rate, or an indicator thereof, that is fed back from the receiver. This function may be implemented using a lookup table given an indicator of the rate that is fed back, and is typically implemented by rate controller 119. Those of ordinary skill in the art will readily appreciate how to arrange such a function given the particular details of the system being implemented, e.g., the channel statistics, the number of substreams employed, and the like. Typically, the lower the rate of data transmission the lower the number of symbols in the constellation for transmitting data at that rate.

[0014] In accordance with an aspect of the invention, each of gain multipliers 113 applies to the mapped data substream that it receives the gain that was indicated in the feedback from the receiver. In an exemplary embodiment of the invention, the better the particular channel that corresponds to one of gain multipliers 113 the greater the gain that is applied, e.g., in accordance with the principles of waterfilling.

[0015] Each of optional upconverters 115 performs conventional upconverting functionality. In the case of a radio-based system each of upconverters 115 generates a radio frequency signal by modulating a carrier waveform using the gain regulated mapped data substream it receives as an input. Each resulting modulated signal may be supplied to the respective one of optional transmit antennas 117 that may be coupled to each of upconverters 115.

[0016] Rate controller 119 receives the rates, or indicators thereof, via feedback from receiver 103 and derives from the received information the code rate and the constellation size for each substream. Each code rate, or an indicator thereof, is then supplied to the appropriate encoder and the constellation to employ, or an indicator thereof, is supplied to each symbol mapper, in accordance with an aspect of the invention. Thus, rate controller 119 may implement a mapping function to determine the code rate and constellation from the information fed back from receiver 103.

[0017] Gain controller 121 receives the gains, or indicators thereof, via feedback from receiver 103 and derives from the received information the gain to be used

for each substream by the associated one of gain multipliers 113. Note that there is a direct relationship between power and gain. More specifically, power is converted to gain by taking the square root of the power. Thus, power may be an indicator for gain, and vice-versa. If power information is received via feedback, it may easily be converted into the appropriate gain.

[0018] Note that the functionality of rate controller 119 may be incorporated into encoders 107 and symbol mappers 111. Similarly, the functionality of gain controller 121 may be incorporated into gain multipliers 113.

[0019] Each of optional receive antennas 131 receives a signal from each of optional transmit antennas 117. The signals received at each antenna are converted to baseband by the one of optional downconverters 133 to which it is coupled. The resulting baseband signals are fed into channel estimator 135.

[0020] Channel estimator 135 develops an estimate of the channels for each transmit and receive pair. Thus, for N transmit antennas and M receive antennas there are NxM channels. The estimates for each of the channels are collectively arranged into an NxM matrix of the overall channel estimate $\hat{H}$. Additionally, channel estimator 135 develops an estimate of the noise power in the channel, $\hat{\sigma}^2$.

[0021] $\hat{H}$ and $\hat{\sigma}^2$ are supplied to power/rate calculator 137 which, in accordance with the principles of the invention, calculates the rates R and powers P—which, as noted above, correspond directly to gains and are used by transmitter 101 in the form of gains—or indicators thereof, that transmitter 101 should use for each data substream produced by demultiplexer 105. The rates and powers are supplied to transmitter 101 using a feedback channel.

[0022] The processes by which rates and powers are assigned by power/rate calculator 137 is up to the implementor. Those of ordinary skill in the art will be able to develop their own processes given the discussion and examples hereinbelow. In particular, the goal of the process is to assign rates and powers to maximize the total channel capacity. Toward this end, typically, those channels that are of a better quality will be assigned higher rates and greater power.

[0023] Note that receiver 103 does not show a decoder and a deinterleaver. This is because, although a decoder and a deinterleaver are necessary for a complete receiver—to reverse the complementary functions performed in the transmitter prior to supplying data as an output of receiver 103—they are not required for the data streams that are supplied to channel estimator 135, and so that are not shown for the sake of clarity of exposition.

[0024] FIG. 2 shows an exemplary process for determining rates and powers for a system with N transmit substreams and M receive branches, in accordance with an aspect of the invention. In one embodiment of the invention, the process of FIG. 2 may be constantly running. However, the values determined by the process are only fed back when there is a significant deviation from the values that were previously fed back. In another embodiment of the invention, the process may begin to run only when channel estimator 135 (FIG. 1) determines that the channel has changed by an amount sufficient to warrant the running of the process. For example, when the norm of the difference of $\hat{H}$ at the last time rates and powers were determined and $\hat{H}$ at the current time is greater than a prescribed threshold. The process of FIG. 2 is performed by power/rate calculator 137 (FIG. 1)

[0025] The process is entered in step 200 (FIG. 2) when it is determined that the rates and powers are to be computed. In step 201 several variables are initialized. In particular, a counter, n, is initialized to the value of N and the value of variable $P_{remaining}$ is initialized to P_T, which is the total transmit power available in the system in which the process of FIG. 2 is being employed. Next, in step 203, an initial power allocation of $P_{remaining}/n$, denoted as P_n, is assigned to the n^{th} substream. The value of $R_n = \log_2(1+P_n\mathbf{h}_n^H(\mathbf{H}_{n+1:N}\mathbf{P}_{n+1:N}\mathbf{H}_{n+1:N}^H+\mathrm{I})^{-1}\mathbf{h}_n)$ (n = 1, ..., N) is calculated in step 205, where

$$\mathbf{h}_n = \left[h_{1,n} \dots h_{M,n}\right]^T$$

is the complex M-dimensional vector for the n^{th} transmit substream;

$h_{m,n}$ is the complex channel coefficient from the n^{th} transmit substream to the m^{th} receive branch, with m = 1 ... M;

superscript T indicates the matrix transpose operation;

$\mathbf{H}_{n+1:N} = [\mathbf{h}_{n+1} \dots \mathbf{h}_N]$ is an M-by-(N-n) matrix;

$\mathbf{P}_{n+1:N} = diag(P_{n+1} \dots P_N)$ is the diagonal (N-m) matrix of assigned powers;

superscript H indicates the Hermitian transpose operation;

superscript -1 denotes the matrix inverse; and

I is the identity matrix of size M x M.

[0026] In step 207, the value of R_n is quantized to the nearest step size, e.g., the nearest integer, the nearest integer which is a multiple of a selected integer, or the nearest multiple of a selected value, or the like. The quantized value of R_n is denoted as $\bar{R}_n$.

[0027] The power of the current substream n is recalculated in step 209. This may be performed by calculating

$$\bar{P}_n = \frac{2^{\bar{R}_n} - 1}{\mathrm{h}_n^H(\mathrm{H}_{n+1:N}\mathrm{P}_{n+1:N}\mathrm{H}_{n+1:N}^H + \mathrm{I})^{-1}\mathrm{h}_n}$$

, where $\bar{P}_n$ indicates a recalculated power. Conditional branch point 211 tests to determine if $P_{remaining}-\bar{P}_n>0$.

This test determines if the amount of power remaining is greater than the amount of power that is allocated for substream n, i.e., the power allocation can actually be performed since there is enough power remaining to support it. If the test result in step 211 is YES, indicating that the power allocation can actually be performed, control passes to step 213, in which the counter n is decremented and $P_{remaining}$ is set to $P_{remaining} - \bar{P}_n$.

[0028] Conditional branch point 215 tests to determine if n=0, i.e., have all the substreams been processed. If the test result in step 215 is NO, control passes back to step 203 and the process continues as described above. If the test result in step 215 is YES, control passes to step 217 and the process exits.

[0029] If the test result in step 211 is NO, indicating that the power allocation cannot actually be performed, control passes to step 219 in which $\bar{R}_n$ is set equal to value of R_n quantized down to the nearest step size value that is less than R_n, e.g., the nearest integer lower than R_n, the nearest integer which is a multiple of a selected integer and is lower than R_n, or the nearest multiple of a selected value and is lower than R_n, or the like. This will result in a lower value of $\bar{P}_n$. Control then passes back to step 209 and the process continues as described above.

[0030] Once the process of FIG. 2 has completed the rates and powers that were generated may be supplied via a feedback path for use in a transmitter. Alternatively, the rates and powers may be encoded so that they are represented by indicators which may be interpreted by the transmitter to determine the appropriate rates and powers—and hence gains. Also, as noted above, the powers may be converted into gains in the receiver, and the gain information directly, or encoded representations thereof, may be supplied via the feedback path for use in the transmitter. Furthermore, in accordance with an aspect of the invention, only information about the rates, or only information about the powers—and hence gains—may be fed back to achieve an improvement over the prior art, although feeding back both results in better performance.

Claims

1. A transmitter of a multiple-input multiple-output (MIMO) system for transmitting a plurality of data substreams derived from a data stream, comprising:

   means for receiving as feedback an indicator of a rate and a power for each data substream; and
   means for applying to each respective data substream a rate control and a power control corresponding to said indicator of a rate and a power received for said data substream.

2. The invention as defined in claim 1 wherein said indicator of a rate and a power includes an independent denotation of said rate and an independent denotation of said power.

3. The invention as defined in claim 1 wherein said indicator of a rate and a power is said rate and said power.

4. The invention as defined in claim 1 wherein said indicator of a rate and a power is a value which is directly a function of said rate and said power indicated by said indicator.

5. The invention as defined in claim 1 wherein each rate and power are determined by a receiver as a function of channel estimates.

6. A receiver of a multiple-input multiple-output (MIMO) system for receiving a plurality of data substreams derived from a data stream, comprising:

   means for estimating channel characteristics; and
   means for deriving an indicator of a rate and an indicator of a power for each data substream as a function of estimated channel characteristics developed by said means for estimating; and
   means for transmitting said indicator of said rate and said indicator of said power as feedback to a transmitter.

7. A transmitter of a multiple-input multiple-output (MIMO) system for transmitting a plurality of data substreams derived from an overall data stream, comprising:

   a plurality of encoders, one encoder for each of said data substreams, each of said encoders being responsive to rate information received as feedback from a receiver of said substreams transmitted by said transmitter; and
   a plurality of symbol mappers, each of said symbol mappers coupled to receive a respective encoded version of one of said data substreams, and each of said symbol mappers also being responsive to said rate information.

8. The invention as defined in claim 7 further comprising a plurality of gain multipliers responsive to power information received as part of said feedback from said receiver.

9. The invention as defined in claim 7 further comprising a plurality of interleavers, each respective one of said interleavers being coupled between a respective one of said encoders and a respective one of said symbol mappers.

10. The invention as defined in claim 7 wherein each of said symbol mappers is associated with a respective one of said encoders for a respective one of said data substreams.

11. The invention as defined in claim 7 wherein each of said symbol mappers is associated with a respective one of said encoders along a transmit path for a respective one of said data substreams, and for each transmit path its associated encoder and symbol mapper receive the same rate from said rate information.

12. The invention as defined in claim 7 further comprising a rate controller for supplying to each of said symbol mappers and said encoders a rate to be used in response to said rate information.

13. The invention as defined in claim 7 further comprising a rate controller for supplying to each of said symbol mappers and said encoders a rate to be used in response to said rate information as a function of said rate information which is received in an encoded format.

14. The invention as defined in claim 7 further comprising a rate controller for supplying to each of said symbol mappers and said encoders a rate to be used in response to said rate information.

15. A transmitter of a multiple-input multiple-output (MIMO) system for transmitting a plurality of data substreams derived from an overall data stream, comprising:

a plurality of gain appliers, one gain applier for each of said data substreams, each of said gain appliers being responsive to power information received as feedback from a receiver of said substreams transmitted by said transmitter.

16. The invention as defined in claim 15 wherein said gain applier is a multiplier.

17. The invention as defined in claim 15 further comprising a gain controller for supplying to each of said gain appliers a power to be used in response to said power information.

18. The invention as defined in claim 15 further comprising a gain controller for supplying to each of said gain appliers a power to be used as a function said power information which is received in an encoded format.

19. A receiver of a multiple-input multiple-output (MIMO) system for receiving a plurality of data substreams derived from an overall data stream, comprising:

a channel estimator for developing (i) an overall channel estimate from said received plurality of data substreams and (ii) an estimate of noise power in said overall channel; and
a power calculator for calculating powers, one for each respective one of said substreams, to be used by a transmitter of said substreams.

20. A receiver of a multiple-input multiple-output (MIMO) system for receiving a plurality of data substreams derived from an overall data stream, comprising:

a channel estimator for developing (i) an overall channel estimate from said received plurality of data substreams and (ii) an estimate of noise power in said overall channel; and
a rate calculator for calculating rates, one for each respective one of said substreams, to be used by a transmitter of said substreams.

21. A method for use in processing for transmission in a multiple-input multiple-output (MIMO) system a plurality of data substreams derived from an overall data stream, the method comprising the steps of:

encoding each of said data substreams as a function of each indicator of a respective rate received as feedback from a receiver of said MIMO system for each of said data substreams to produce encoded data substreams; and
mapping each of said encoded data substreams after they are each respectively interleaved, said mapping for each of said encoded interleaved data stream being performed using a constellation selected as a function of each of said indicators of said respective rate received as feedback from a receiver of said MIMO system for each of said data substreams to produce encoded interleaved mapped data substreams.

22. The invention as defined in claim 21 further comprising the step of controlling the power of each of said encoded interleaved mapped data substreams as a function of respective indicators received as feedback from said receiver of said MIMO system for power control of each of said encoded interleaved mapped data substreams.

23. The invention as defined in claim 22 wherein there are N transmit paths, one for each data substream and M receive paths, so that there are NxM channels, said indicators and said rates are developed in said receiver by the steps of:

developing overall channel estimate $\hat{\mathbf{H}}$, which is an NxM matrix;
determine a noise power in the channel, $\hat{\sigma}^2$.
initializing counter n to the value of N;
initializing variable $P_{remaining}$ to the total transmit power available in the system P_T;
assigning an initial power allocation of the n^{th} substream P_n to a value of $P_{remaining}/n$;
computing $R_n = \log_2(1+P_n\mathbf{h}_n^H(\mathbf{H}_{n+1:N}\mathbf{P}_{n+1:N}\mathbf{H}^H_{n+1:N}+\mathrm{I})^{-1}\mathbf{h}_n)$ $(n = 1, ..., N)$ where

R_n is a rate to use for the for the n^{th} transmit substream;

$$\mathbf{h}_n = \left[h_{1,n} \ldots h_{M,n}\right]^T$$

is the complex M-dimensional vector for the n^{th} transmit substream;
$h_{m,n}$ is the complex channel coefficient from the n^{th} transmit substream to the m^{th} receive branch, with $m = 1 \ldots M$;
superscript T indicates the matrix transpose operation;
$\mathbf{H}_{n+1:N} = [\mathbf{h}_{n+1} \ldots \mathbf{h}_N]$ is an M-by-$(N$-$n)$ matrix;
$\mathbf{P}_{n+1:N} = diag(P_{n+1} \ldots P_N)$ is the diagonal $(N$-$m)$ matrix of assigned powers;
superscript H indicates the Hermitian transpose operation;
superscript -1 denotes the matrix inverse; and
I is the identity matrix of size M x M;

quantizing R_n to the nearest step size to develop $\bar{R}_n$; and
recalculating the power of current substream n by calculating

$$\bar{P}_n = \frac{2^{\bar{R}_n} - 1}{\mathrm{h}_n^H(\mathrm{H}_{n+1:N}\mathrm{P}_{n+1:N}\mathrm{H}^H_{n+1:N} + \mathrm{I})^{-1}\mathrm{h}_n}$$

, where $\bar{P}_n$ indicates a recalculated power.

24. The invention as defined in claim 22 further comprising the following steps when $P_{remaining} - \bar{P}_n > 0$

decrementing n;
setting $P_{remaining}$ to $P_{remaining} - \bar{P}_n$; and
when n is not equal to zero, repeating said assigning, computing, quantizing, and recalculating steps.

25. The invention as defined in claim 22 further comprising the following steps when $P_{remaining} - \bar{P}_n \leq 0$

setting $\bar{R}_n$ equal to value of R_n quantized down to the nearest step size value that is less than R_n; and
repeating said recalculating step.

26. A method for use in a receiver of a multiple-input multiple-output (MIMO) system, in which there are N transmit paths, one for each data substream to be transmitted and M receive paths, so that there are NxM channels in an overall channel, the method comprising the steps of:

developing overall channel estimate $\hat{\mathbf{H}}$, which is an NxM matrix;
determine a noise power in the channel, $\hat{\sigma}^2$.
initializing counter n to the value of N;
initializing variable $P_{remaining}$ to the total transmit power available in the system P_T.
assigning an initial power allocation of the n^{th} substream P_n to a value of $P_{remaining}/n$;
computing $R_n = \log_2(1 + P_n\mathbf{h}_n^H(\mathbf{H}_{n+1:N}\mathbf{P}_{n+1:N}\mathbf{H}^H_{n+1:N} + \mathrm{I})^{-1}\mathbf{h}_n)$ $(n = 1, ..., N)$ where

R_n is a rate to use for the for the n^{th} transmit substream;
$\mathbf{h}_n = [h_{1,n} \ldots h_{M,n}]^T$ is the complex M-dimensional vector for the n^{th} transmit substream;
$h_{m,n}$ is the complex channel coefficient from the n^{th} transmit substream to the m^{th} receive branch, with $m = 1 \ldots M$;
superscript T indicates the matrix transpose operation;
$\mathbf{H}_{n+1:N}=[\mathbf{h}_{n+1} \ldots \mathbf{h}_N]$ is an M-by-$(N$-$n)$ matrix;
$\mathbf{P}_{n+1:N} = diag(P_{n+1} \ldots P_N)$ is the diagonal $(N$-$m)$ matrix of assigned powers;
superscript H indicates the Hermitian transpose operation;
superscript -1 denotes the matrix inverse; and
I is the identity matrix of size M x M;

quantizing R_n to the nearest step size to develop $\bar{R}_n$; and
recalculating the power of current substream n by calculating

$$\bar{P}_n = \frac{2^{\bar{R}_n} - 1}{\mathrm{h}_n^H(\mathrm{H}_{n+1:N}\mathrm{P}_{n+1:N}\mathrm{H}^H_{n+1:N} + \mathrm{I})^{-1}\mathrm{h}_n}$$

, where $\bar{P}_n$ indicates a recalculated power.

27. The invention as defined in claim 26 further comprising the following steps when $P_{remaining} - \bar{P}_n > 0$

decrementing n;
setting $P_{remaining}$ to $P_{remaining} - \bar{P}_n$; and
when n is not equal to zero, repeating said assigning, computing, quantizing, and recalculating steps.

28. The invention as defined in claim 26 further comprising the following steps when $P_{remaining} - \bar{P}_n \leq 0$

setting $\bar{R}_n$ equal to value of R_n quantized down to the nearest step size value that is less than R_n; and
repeating said recalculating step.

FIG. 1

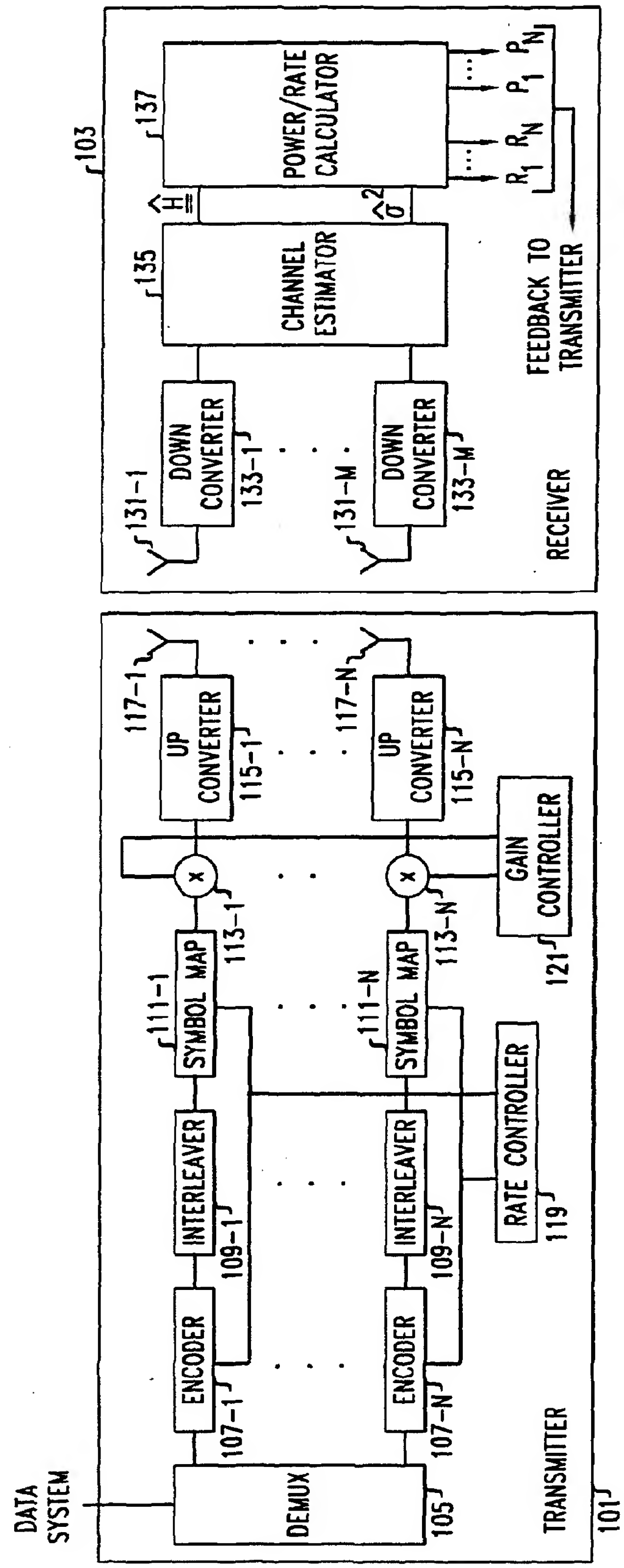
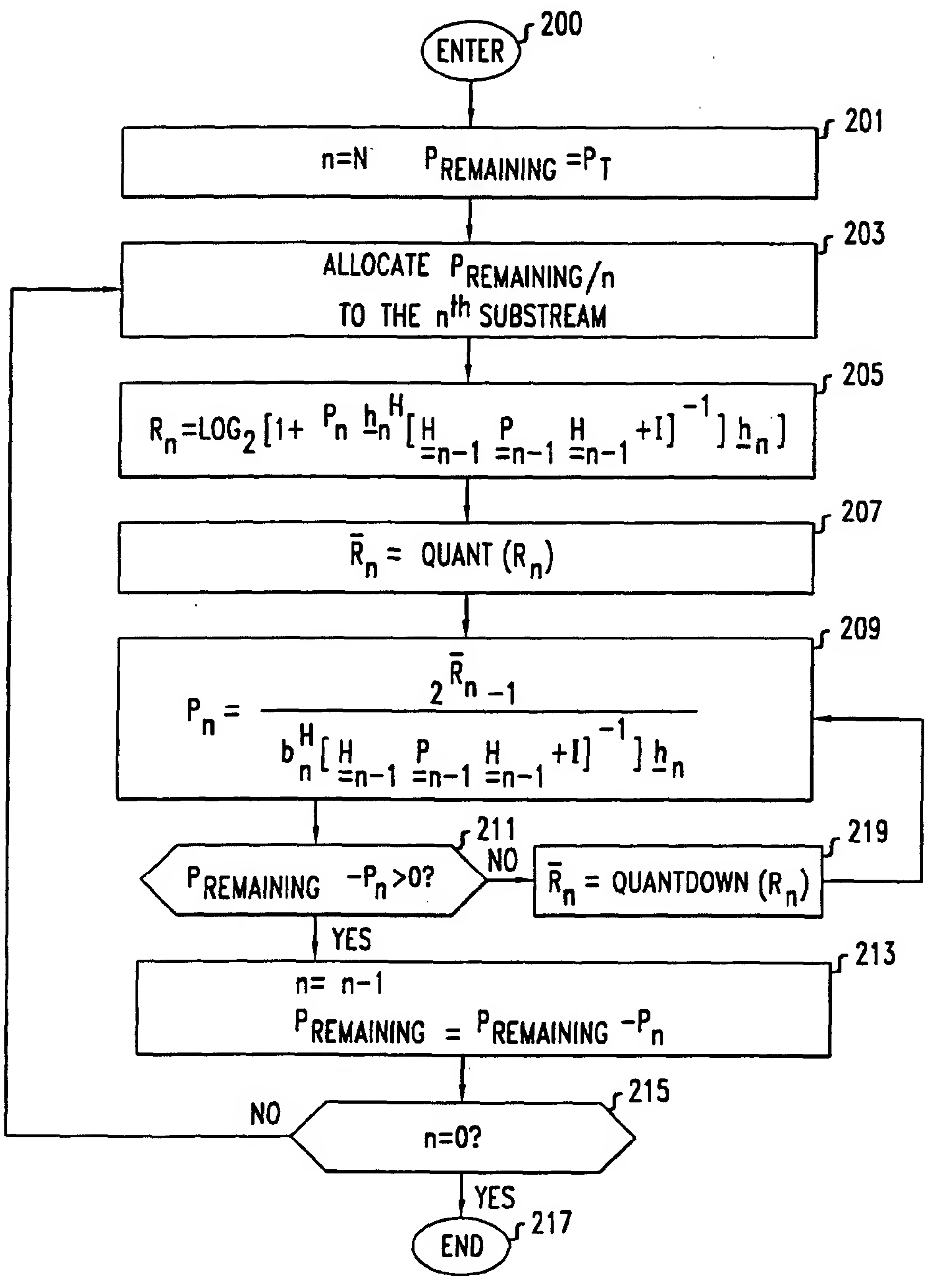

DATA
SYSTEM
DEMUX
105
ENCODER
107-1
107-N
INTERLEAVER
109-1
109-N
SYMBOL MAP
111-1
111-N
113-1
113-N
UP
CONVERTER
115-1
115-N
117-1
117-N
RATE CONTROLLER
119
GAIN
CONTROLLER
121
TRANSMITTER
101
103
131-1
131-M
DOWN
CONVERTER
133-1
133-M
CHANNEL
ESTIMATOR
135
POWER/RATE
CALCULATOR
137
R1 RN P1 PN
FEEDBACK TO
TRANSMITTER
RECEIVER

FIG. 2


ENTER
200
n=N $P_{REMAINING}=P_T$
201
ALLOCATE $P_{REMAINING}/n$
TO THE n^{th} SUBSTREAM
203
$R_n=LOG_2[1+P_n \underline{h}_n^H[\underline{\underline{H}}_{n-1}\underline{\underline{P}}_{n-1}\underline{\underline{H}}_{n-1}+I]^{-1}]\underline{h}_n]$
205
$\bar{R}_n=$ QUANT (R_n)
207
$P_n=\frac{2^{\bar{R}_n}-1}{b_n^H[\underline{\underline{H}}_{n-1}\underline{\underline{P}}_{n-1}\underline{\underline{H}}_{n-1}+I]^{-1}]\underline{h}_n}$
209
$P_{REMAINING}-P_n>0?$
211
NO
$\bar{R}_n=$ QUANTDOWN (R_n)
219
YES
n= n−1
$P_{REMAINING}=P_{REMAINING}-P_n$
213
NO
n=0?
215
YES
END
217

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 30 4722

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
Y	WO 00 13362 A (HALFMANN RUEDIGER ;EICHINGER JOSEF (DE); LIEGL WERNER (DE); SIEMEN) 9 March 2000 (2000-03-09) * abstract * * claims 1-18 * * figures 2,3 * ---	1-28	H04L1/00 H04B7/06
Y	KOUSA M A ET AL: "MULTICHANNEL ADAPTIVE SYSTEM" IEE PROCEEDINGS I. SOLID- STATE & ELECTRON DEVICES, INSTITUTION OF ELECTRICAL ENGINEERS. STEVENAGE, GB, vol. 140, no. 5, PART 1, 1 October 1993 (1993-10-01), pages 357-364, XP000403498 ISSN: 0956-3776 * the whole document * ---	1-28	
A	US 6 084 919 A (CHUPRUN JEFFERY SCOTT ET AL) 4 July 2000 (2000-07-04) * abstract * * column 8, line 10-25 * * column 8, line 56-65 * * figures 1,2 * ---	1-28	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04L H04B
A	KNOPP R ET AL: "Power control schemes for TDD systems with multiple transmit and receive antennas" GLOBAL TELECOMMUNICATIONS CONFERENCE - GLOBECOM '99, 1999, pages 2326-2330, XP010373369 * the whole document * -----	1-28	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	18 March 2002	Yang, Y

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 01 30 4722

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report.
The members are as contained in the European Patent Office EDP file on
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18-03-2002

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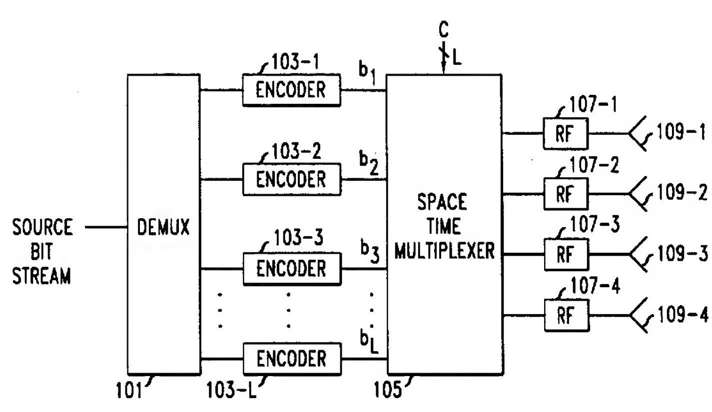
(54) Open-loop diversity technique for systems employing four transmitter antennas

(57) When using four transmit antennas, conventional channel coding is employed for a decoupled space-time coding approach for each of a number, L, of data substreams derived from the overall source bit stream. The symbols of the data substreams, after any encoding, are processed and the resulting derivatives of the encoded data substreams, which includes at least the complex conjugate of one of the encoded symbols, are grouped to form four transmit time sequences each one spanning L symbol periods which form a transmission matrix **B**. Each row of the matrix corresponds to an antenna, and the elements of each row represent the samples of a temporal sequence that is emitted by the antenna in L symbol periods. When L=4 the matrix **B** can be arranged as follows:

$$\begin{array}{r c} & \begin{array}{cccc} T_1 & T_2 & T_3 & T_4 \end{array} \\ \begin{array}{r} \textit{Antenna } 1 \\ \textit{Antenna } 2 \\ \textit{Antenna } 3 \\ \textit{Antenna } 4 \end{array} & \begin{bmatrix} b_1 & b_2^* & b_3 & b_4^* \\ b_2 & -b_1^* & -b_4 & b_3^* \\ b_3 & b_4^* & -b_1 & -b_2^* \\ b_4 & -b_3^* & b_2 & -b_1^* \end{bmatrix} \end{array}$$

where b_1, b_2, b_3, and b_4 are the encoded symbols from the data substreams and * indicates complex conjugate.

FIG. 1



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Description

Technical Field

[0001] This invention relates to the art of wireless communications, and more particularly, to wireless communication systems using multiple antennas at the transmitter and one or more antennas at the receiver.

Background of the Invention

[0002] It is known in the art that systems using multiple antennas at the transmitter and one or more antennas at the receiver can achieve dramatically improved capacity, i.e., the maximum bits/second/hertz with error free communication, as compared to single transmitter antenna systems. It is also known in the art that if a channel estimate, or channel statistics based on the channel estimate, are fed back to the transmitter, then the throughput of the channel can be improved with respect to an identically configured system but without feedback. However, because in systems with multiple transmit antennas the overall channel is actually made up of multiple channels, with one channel for each transmit and receive antenna pairing, such feedback requires considerable bandwidth, and it is undesirable to dedicate so much bandwidth to feedback. Also, for fast changing wireless channels, the feedback may not arrive at a fast enough rate in order to be useful.

[0003] In order to achieve the maximum open loop capacity of a multiple transmit antenna system, it is necessary to employ channel coding. The state of the art of channel coding, e.g., turbo codes, trellis codes and the like, is typically spatially one dimensional, i.e., they are designed for only a single transmit antenna. Generally, it is not immediately obvious how one could employ such coding in a spatially multi-dimensional, e.g., multiple transmit antenna, system. In the case of two transmit antennas and one receive antenna, it was recognized that each of the data substreams could be independently coded using known channel coding techniques to achieve maximum capacity if so-called "space-time spreading" was used. See for example, United States Patent Application Serial No. 09/285578 entitled Space-Time Spreading Method Of CDMA Wireless Communication.

Summary of the Invention

[0004] We have recognized that it is possible to employ conventional channel coding when using four transmit antennas, by using a decoupled space-time coding approach for each of a number, L, of data substreams that are derived from the overall source bit stream. To this end symbols, which are samples of the data substreams after encoding, if any, are processed so that the resulting derivatives thereof, which includes at least the complex conjugate of one of the symbols, are grouped to form four transmit time sequences each spanning L symbol periods. The four time sequences may be viewed as forming a transmission matrix **B**. Each row of the matrix corresponds to a transmit element, e.g., an antenna, and the elements of each row represent the samples of a temporal sequence that is emitted by the antenna in L symbol periods. In one embodiment of the invention, the number of data substreams L is equal to four, the number of antennas. The overall symbol rate for each antenna is the same as the symbol rate of the original overall source bit stream. In accordance with the principles of the invention, the matrix **B** is arranged as follows:

$$
\begin{array}{r c}
 & \begin{array}{cccc} T_1 & T_2 & T_3 & T_4 \end{array} \\
\begin{array}{r} \textit{Antenna } 1 \\ \textit{Antenna } 2 \\ \textit{Antenna } 3 \\ \textit{Antenna } 4 \end{array} &
\begin{bmatrix}
b_1 & b_2^* & b_3 & b_4^* \\
b_2 & -b_1^* & -b_4 & b_3^* \\
b_3 & b_4^* & -b_1 & -b_2^* \\
b_4 & -b_3^* & b_2 & -b_1^*
\end{bmatrix}
\end{array}
$$

where b_1, b_2, b_3, and b_4 are the symbols, which may be encoded samples, derived from data substreams 1, 2, 3, and 4, respectively, and * indicates complex conjugate. As indicated, the rows of the matrix represent the different antennas, while the columns represent different symbol periods (T_i, i=1 .. 4).

[0005] This approach may be used in systems that employ direct sequence spreading, e.g., code division multiple access (CDMA) systems. In such systems, each entry of the matrix is modulated, i.e., multiplied, by an orthogonal spreading code sequence represented by one of horizontal vectors $\bar{c}_l$, l=1 .., L, each of which spans 1 symbol period and contains N chips, where N is the spreading gain.

[0006] In a receiver, the received signal is first applied to a bank of L correlators, where despreading, in the case of a spread spectrum signal, or otherwise the simpler version thereof, temporal subsampling, is performed, producing L

preprocessed sequences. The preprocessed sequences are further processed, either jointly or individually, to ultimately develop a reconstructed version of the source bit stream.

[0007] Advantageously, using this methodology, the samples of the data substreams may be encoded, as indicated, prior to being incorporated into the matrix using conventional coding techniques, e.g., turbo coding, and the advantages of such coding may be exploited at the receiver. Further advantageously, the coding of each of the data substreams is independent of, i.e., decoupled from, the coding of any of the other data substreams.

Brief Description of the Drawing

[0008] In the drawing:

FIG. 1 shows an exemplary transmitter arranged in accordance with the principles of the invention; and
FIG. 2 shows an exemplary embodiment of a receiver arranged in accordance with the principles of the invention.

Detailed Description

[0009] The following merely illustrates the principles of the invention. It will thus be appreciated that those skilled in the art will be able to devise various arrangements which, although not explicitly described or shown herein, embody the principles of the invention and are included within its spirit and scope. Furthermore, all examples and conditional language recited herein are principally intended expressly to be only for pedagogical purposes to aid the reader in understanding the principles of the invention and the concepts contributed by the inventor(s) to furthering the art, and are to be construed as being without limitation to such specifically recited examples and conditions. Moreover, all statements herein reciting principles, aspects, and embodiments of the invention, as well as specific examples thereof, are intended to encompass both structural and functional equivalents thereof. Additionally, it is intended that such equivalents include both currently known equivalents as well as equivalents developed in the future, i.e., any elements developed that perform the same function, regardless of structure.

[0010] Thus, for example, it will be appreciated by those skilled in the art that any block diagrams herein represent conceptual views of illustrative circuitry embodying the principles of the invention. Similarly, it will be appreciated that any flow charts, flow diagrams, state transition diagrams, pseudocode, and the like represent various processes which may be substantially represented in computer readable medium and so executed by a computer or processor, whether or not such computer or processor is explicitly shown.

[0011] The functions of the various elements shown in the FIGs., including functional blocks labeled as "processors", may be provided through the use of dedicated hardware as well as hardware capable of executing software in association with appropriate software. When provided by a processor, the functions may be provided by a single dedicated processor, by a single shared processor, or by a plurality of individual processors, some of which may be shared. Moreover, explicit use of the term "processor" or "controller" should not be construed to refer exclusively to hardware capable of executing software, and may implicitly include, without limitation, digital signal processor (DSP) hardware, read-only memory (ROM) for storing software, random access memory (RAM), and non-volatile storage. Other hardware, conventional and/or custom, may also be included. Similarly, any switches shown in the FIGS. are conceptual only. Their function may be carried out through the operation of program logic, through dedicated logic, through the interaction of program control and dedicated logic, or even manually, the particular technique being selectable by the implementor as more specifically understood from the context.

[0012] In the claims hereof any element expressed as a means for performing a specified function is intended to encompass any way of performing that function including, for example, a) a combination of circuit elements which performs that function or b) software in any form, including, therefore, firmware, microcode or the like, combined with appropriate circuitry for executing that software to perform the function. The invention as defined by such claims resides in the fact that the functionalities provided by the various recited means are combined and brought together in the manner which the claims call for. Applicant thus regards any means which can provide those functionalities as equivalent as those shown herein.

[0013] Unless otherwise explicitly specified herein, the drawings are not drawn to scale.

[0014] FIG. 1 shows an exemplary transmitter arranged in accordance with the principles of the invention. The transmitter of FIG. 1 uses a decoupled space-time coding approach for each of a number, L, of data substreams that are derived from a source bit stream. The L data substreams are processed so that they may be transmitted via four transmit antennas. Advantageously, each data substream employ conventional channel coding.

[0015] FIG. 1 shows a) demultiplexer (DEMUX) 101; b) encoders 103, including encoders 103-1 through 103-L; c) space time multiplexer 105; d) radio frequency (RF) units 107, including RF units 107-1 through 107-4; and e) antennas 109, including antennas 109-1 through 109-4.

[0016] Demultiplexer 101 divides the source bit stream it receives as an input into L data substreams. Each of the

L data substreams supplied as an output by demultiplexer 101 is optionally encoded by a respective one of optional encoders 103 to produce encoded data substreams. Advantageously, encoders 103 may employ conventional channel coding. such as turbo coding. Encoders 103 may also perform digital modulation, e.g., mapping the samples to a discrete alphabet prior to doing the actual encoding. The encoded data substreams are supplied to space time multiplexer 105.

[0017] Each sample supplied as an output by one of encoders 103, or by demultiplexer 101 in the event encoders 103 are omitted, is referred to herein as a symbol. The time duration of a symbol is referred to as a symbol period.

[0018] Every symbol period, space time multiplexer 105, processes the symbols of each of the encoded data substreams supplied by encoders 103 so as to form four transmit time sequences, each time sequence spanning at least L symbol periods. Each of the symbols is processed to develop its complex conjugate. In one embodiment of the invention, where L=4, the four symbols, their complex conjugates, the negative of the symbols and the negative of the complex conjugates are arranged by space time multiplexer 105 to form a matrix **B.** Each row of matrix **B** corresponds to one of antennas 109. More specifically, the elements of each row represent the samples of a temporal sequence that is emitted by the corresponding one of antennas 109 in L symbol periods, unless direct sequence spreading is used, as described further hereinbelow.

[0019] In such an embodiment of the invention, the matrix **B** is arranged as follows:

$$\begin{array}{r c} & \begin{array}{cccc} \mathrm{T1} & \mathrm{T2} & \mathrm{T3} & \mathrm{T4} \end{array} \\ \begin{array}{r} \textit{Antenna}\ 1 \\ \textit{Antenna}\ 2 \\ \textit{Antenna}\ 3 \\ \textit{Antenna}\ 4 \end{array} & \begin{bmatrix} b_1 & b_2^* & b_3 & b_4^* \\ b_2 & -b_1^* & -b_4 & b_3^* \\ b_3 & b_4^* & -b_1 & -b_2^* \\ b_4 & -b_3^* & b_2 & -b_1^* \end{bmatrix} \end{array}$$

where b_1, b_2, b_3, and b_4 are the symbols from the encoded data substreams supplied as outputs from encoders 103-1, 103-2, 103-3, and 103-4, respectively, and * indicates complex conjugate. As indicated, the rows of matrix **B** represent the different antennas, while the columns represent different symbol periods (Ti, i=1 .. 4).

[0020] In the event that the inventive technique is employed in a system that employs direct sequence spreading, e.g., code division multiple access (CDMA) system, space time multiplexer 105 further multiplies each element of the l^{th} column, l=1 ... L, of an unspread matrix **B** by a spreading code sequence represented by $\bar{c}_l$, which spans 1 symbol period arid contains N chips, where N is the spreading gain. The set $\bar{c}_l$, l=1 ... L, may be orthogonal. Thus, the spread matrix **B,** which is transmitted, becomes

$$\begin{bmatrix} b_1\bar{c}_1 & b_2^*\bar{c}_2 & b_3\bar{c}_3 & b_4^*\bar{c}_4 \\ b_2\bar{c}_1 & -b_1^*\bar{c}_2 & -b_4\bar{c}_3 & b_3^*\bar{c}_4 \\ b_3\bar{c}_1 & b_4^*\bar{c}_2 & -b_1\bar{c}_3 & -b_2^*\bar{c}_4 \\ b_4\bar{c}_1 & -b_3^*\bar{c}_2 & b_2\bar{c}_3 & -b_1^*\bar{c}_4 \end{bmatrix}$$

[0021] Each of the time sequences developed by space time multiplexer 105 are supplied as an input to a respective one of radio frequency (RF) units 107, each of which performs all the necessary processing to convert its respective input from baseband to a radio frequency modulated signal. Each of the radio frequency modulated signal developed by each of RF units 107 is supplied to a respective one of antennas 109 from which it is transmitted.

[0022] Note that although antennas are shown in FIG. 1, any form of transmit element may be employed, e.g., a light source. Further note that although radio frequency units are shown in FIG. 1, in other embodiments of the invention, e.g., those using light for communicating the transmitted signal, different modulators may be employed

[0023] FIG. 2 shows an exemplary embodiment of a receiver arranged in accordance with the principles of the invention. FIG. 2 shows a) antenna 201; b) radio frequency (RF) units 203; c) demultiplexer 202; d) correlator 205-1 through 205-L; e) selective conjugator 206, f) matrix multiplier 207; g) baseband signal processing unit 209; h) optional decoders 211, including decoders 211-1 through 211-L; and i) multiplexer 213.

[0024] Antenna 201 receives the signals transmitted by all of antennas 109 (FIG. 1) and supplies an electrical version thereof to RF unit 203 (FIG. 2). RF unit 203 converts the radio frequency signal supplied to it by antenna 201 to a baseband version thereof.

[0025] Demultiplexer (demux) 202 performs subsampling, i.e., it divides the received baseband signal into L portions

in time, and supplies one portion to each respective one of L outputs to form data substreams which are supplied to optional correlators 205 or directly to respective inputs of selective conjugator 206.

[0026] In the event that the inventive technique is employed in a system that employs direct sequence spreading, e.g., a code division multiple access (CDMA) system, each of optional correlators 205 is supplied with a respective one of an orthogonal spreading code sequence represented as a horizontal vector $\bar{c}_l$, l=1 .., L where each of orthogonal spreading code sequences $\bar{c}_l$, spans 1 symbol period and contains N chips, where N is the spreading gain. Thus, correlators 205 perform despreading which is the inverse of the spreading performed in space time multiplexer 105, and each supplies as an output a despread data substream which is supplied to selective conjugator 206.

[0027] Selective conjugator 206 determines the complex conjugate of any of the outputs d'_i of demultiplexer 202 or optional correlators 205 that is required to ensure that the system is not over parametrized. In other words, there should only be one form of each symbol that is being sought in the system of linear equations that describes the input to matrix multiplier 207. This system of linear equations is generally represented, when the channel is a flat-faded channel and in the absence of noise, as **d=Hb,** where **d** is the vertical vector that is the output of selective conjugator 206, **H** is a matrix of derivatives of channel coefficients h_1, h_2, h_3, and h_4, **b** is a vertical vector formed from b_1... b_L., i.e., if symbol b_i appears in **b**, then b^*_i should not appear. This is necessary in order that the inputs to matrix multiplier 207 be a linear function of only the symbols b_i, i=1 .. L.

[0028] For example, if L=4 and matrix **B** is arranged in the manner described hereinabove, i.e.,

$$\begin{bmatrix} b_1 & b_2^* & b_3 & b_4^* \\ b_2 & -b_1^* & -b_4 & b_3^* \\ b_3 & b_4^* & -b_1 & -b_2^* \\ b_4 & -b_3^* & b_2 & -b_1^* \end{bmatrix}$$

$$\text{then } \mathbf{d}=\begin{bmatrix} d'_1 \\ d'^*_2 \\ d'_3 \\ d'^*_4 \end{bmatrix} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ -h_2^* & h_1^* & -h_4^* & h_3^* \\ -h_3 & h_4 & h_1 & -h_2 \\ -h_4^* & -h_3^* & h_2^* & h_1^* \end{bmatrix} \begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix}.$$

so it is necessary to complex conjugate d'_2 and d'_4 to develop **d=Hb.**

[0029] Matrix multiplier 207 operates on the received vertical vector **d** to produce L match filtered outputs. To this end, matrix multiplier 207 also receives, or derives, an L by L matrix $\mathbf{H}^\dagger$, where † denotes complex conjugate transpose, also known as Hermitian transpose. In an embodiment of the invention where L is 4, **H**, as noted above, is the following matrix

$$\begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ -h_2^* & h_1^* & -h_4^* & h_3^* \\ -h_3 & h_4 & h_1 & -h_2 \\ -h_4^* & -h_3^* & h_2^* & h_1^* \end{bmatrix}$$

where h_i is the complex channel coefficient from the i^{th} transmit antenna to the receiver antenna assuming all the channels are flat faded channels. The matrix $\mathbf{H}^\dagger$, multiplies from the left the L by 1 vertical vector **d** formed by the outputs of correlators 205 to produce a new L by 1 vertical vector **f** whose L entries are the inputs to baseband signal processing unit 209, i.e., $\mathbf{f}=\mathbf{H}^\dagger\mathbf{d}$.

[0030] The L match filtered outputs are supplied to baseband signal processing unit 209 in order to extract therefrom L data subtreams. In various embodiments of the invention, in baseband signal processing unit 209 a specified matrix **W** multiplies from the left vertical vector **f** to produce a new L by 1 vertical vector **r,** so that **r =Wf**. In one embodiment of the invention, a decorrelating process, also known as zero forcing, is employed in which the matrix $\mathbf{W}=\mathbf{K}^{-1}$ is com-

puted, where $\mathbf{K}= \mathbf{H}^{\dagger} \mathbf{H}$. In one embodiment of the invention, $\mathbf{W}=\mathbf{K}^{\dagger}(\mathbf{K} \mathbf{K}^{\dagger}+\lambda \mathbf{K})^{-1}$, where λ is a real scalar. More particularly, λ may be equal to σ_n^2/σ_b^2 where σ_n^2 is the channel noise variance and σ_b^2 is the variance of any of b_i, where i= 1 .. L. Such a receiver is known as a minimum mean squared error (MMSE) receiver.

[0031] In accordance with an aspect of the invention, in performing the multiplication in baseband signal processing unit 209 when L=4 for the foregoing embodiments, it is possible to speed up the calculations with no loss in accuracy by subdividing the matrix multiplication process into two portions, the first portion using the first and third element of vertical vector **f** and the second portion using the fourth and second element of vertical vector **f**, specifically in that order. A particular 2 x 2 matrix that is employed for the multiplication from the left for both portions is derived as follows. First derive 2 x 2 matrix **K'** which corresponds to deleting the second and fourth rows and columns of 4 x 4 matrix **K.** Then, determine 2 x 2 matrix **W'** using the same process by which **W** is derived from **K**, but employing **K'** in lieu of **K** for the respective processing desired, e.g., zero forcing or MMSE.

[0032] In other embodiments of the invention, baseband signal processing unit 209 may employ non-linear processing techniques, such as non-linear multi-user detection, including maximum likelihood multi-user detection and interference cancellation. Such non-linear processing techniques may also employ the partitioning technique described above to reduce computational complexity without sacrificing accuracy.

[0033] Each of the L data substreams is then, optionally, decoded by a respective one of optional decoders 211 to which it is supplied. The decoding performed by decoders 111 advantageously is the inverse of that performed by encoders 103, and as such they may also perform digital demodulation.

[0034] Note that in other embodiments of the invention the decoding process may be combined with the processing of baseband signal processing unit 209. In such embodiments of the invention, it is possible to perform joint decoding, i.e., decoding using information from more than one data stream. Such embodiments of the invention may be especially suitable for the use of non-linear processing techniques in baseband signal processing unit 209. Furthermore, note that decoding may be eliminated entirely if no encoders 103 (FIG. 1) were included in the transmitter.

[0035] The decoded L data substreams are then supplied as an input to multiplexer (MUX) 213 (FIG. 2) which interleaves them in the inverse pattern of DEMUX 101 to reconstruct the source bit stream.

[0036] In other embodiments of the invention, the functionality of matrix multiplier 207 may be absorbed into the processing performed by baseband signal processing unit 209. In yet further embodiments of the invention, the functionality of matrix multiplier 207 may be eliminated.

[0037] Note that the decision functionality which is that part of the process which selects the closest constellation point may be performed in either baseband signal processing unit 209, in decoders 211, or distributed across both at the discretion of the implementor based on the particular decoding selected. Furthermore, the particular algorithm employed to achieve the decision functionality is at discretion of the implementor.

[0038] Advantageously, using this methodology, the samples of the data substreams may be encoded, as indicated, prior to being incorporated into the matrix using conventional coding techniques, e.g., turbo coding, and the advantages of such coding may be exploited at the receiver. Further advantageously, the coding of each of the data substreams is independent of, i.e., decoupled from, the coding of any of the other data substreams.

[0039] In another embodiment of the invention, where L=4, the matrix **B** is arranged as follows:

$$\begin{bmatrix} b_1 & b_2^* & b_3 & b_4^* \\ -b_2 & b_1^* & b_4 & -b_3^* \\ b_3 & b_4^* & -b_1 & -b_2^* \\ -b_4 & b_3^* & -b_2 & b_1^* \end{bmatrix}$$

where b_1, b_2, b_3, and b_4 are the symbols from the encoded data substreams supplied as outputs from encoders 103-1(FIG. 1), 103-2, 103-3, and 103-4, respectively, and * indicates complex conjugate. Similarly, in another embodiment of the invention, the transpose of the immediately preceding matrix B, i.e.,

$$\begin{bmatrix} b_1 & -b_2 & b_3 & -b_4 \\ b_2^* & b_1^* & b_4^* & b_3^* \\ b_3 & b_4 & -b_1 & -b_2 \\ b_4^* & -b_3^* & -b_2^* & b_1^* \end{bmatrix}$$

may be employed.

[0040] In another embodiment of the invention, where L=4, the matrix **B** is arranged as follows:

$$\begin{bmatrix} b_1 & b_2 & b_3 & b_4 \\ b_2^* & -b_1^* & b_4^* & -b_3^* \\ b_3 & -b_4 & -b_1 & b_2 \\ b_4^* & b_3^* & -b_2^* & -b_1^* \end{bmatrix}$$

where b_1, b_2, b_3, and b_4 are the symbols from the encoded data substreams supplied as outputs from encoders 103-1(FIG. 1), 103-2, 103-3, and 103-4, respectively, and * indicates complex conjugate.

[0041] It will be readily recognized by those of ordinary skill in the art that in lieu of starting the process with the set of b_1, b_2, b_3, and b_4, any other set of derivatives of b_1, b_2, b_3, and b_4 may be employed by the implementor, with corresponding changes made throughout the process. Furthermore, the ordering of the derivatives of the substreams is also solely at the discretion of the implementor, again with corresponding changes made throughout the process. Thus, for example, one may select to use the set b_4, b_1^*, $-b_2$, b_3.

[0042] Those of ordinary skill in the art of non-flat faded channels will be able to apply the techniques of the invention for use with non-flat faded channels.

[0043] In another embodiment of the invention, multiple receive elements are employed. The symbols are reconstructed for each of the data substreams developed at each receive element in the manner described hereinabove. They may then be combined to develop an improved estimate of the original symbol. Such combination may be achieved, for example, by averaging values for each corresponding symbol.

Claims

1. A method for use in a transmitter adapted to employ four transmit elements to transmit a source bit stream, the method comprising the steps of:

   dividing said source bit stream into L data substreams, L > 2; and
   grouping derivatives of symbols derived from each of said data substreams to form four transmit time sequences, one sequence for each transmit element, each of said time sequences spanning L symbol periods, at least one of said derivatives of said symbols being a complex conjugate of one of said symbols.

2. The invention as defined in claim 1 wherein L=4 and said time sequences are arranged according to a matrix, each time sequence being a row of said matrix and being transmitted by a respective one of said transmit elements, said matrix being arranged as one of the matrices of the set of matrices consisting of

$$\begin{bmatrix} b_1 & b_2^* & b_3 & b_4^* \\ b_2 & -b_1^* & -b_4 & b_3^* \\ b_3 & b_4^* & -b_1 & -b_2^* \\ b_4 & -b_3^* & b_2 & -b_1^* \end{bmatrix}, \begin{bmatrix} b_1 & b_2^* & b_3 & b_4^* \\ -b_2 & b_1^* & b_4 & -b_3^* \\ b_3 & b_4^* & -b_1 & -b_2^* \\ -b_4 & b_3^* & -b_2 & b_1^* \end{bmatrix}, \begin{bmatrix} b_1 & -b_2 & b_3 & -b_4 \\ b_2^* & b_1^* & b_4^* & b_3^* \\ b_3 & b_4 & -b_1 & -b_2 \\ b_4^* & -b_3^* & -b_2^* & b_1^* \end{bmatrix}, \text{ and}$$

$$\begin{bmatrix} b_1 & b_2 & b_3 & b_4 \\ b_2^* & -b_1^* & b_4^* & -b_3^* \\ b_3 & -b_4 & -b_1 & b_2 \\ b_4^* & b_3^* & -b_2^* & -b_1^* \end{bmatrix}$$

   where:
   b_1, b_2, b_3, and b_4 are said symbol derivatives from data substreams 1, 2, 3, and 4, respectively, and
   * indicates complex conjugate.

3. The invention as defined in claim 1 wherein at least one of said groups of derivatives of said symbols includes derivatives of symbols from more than one of said data substreams.

4. The invention as defined in claim 1 further comprising the step of repeating said dividing and grouping steps.

5. The invention as defined in claim 1 wherein at least one of said derivatives of said symbols is one of the group consisting of: a negative of one of said symbols, a negative of a complex conjugate of one of said symbols, one of said symbols, a symbol developed by encoding at least one sample of at least one of said data substreams, an unencoded sample of at least one of said data substreams.

6. The invention as defined in claim 1 wherein each row of said matrix represents what is transmitted by a respective one of said transmit elements.

7. The invention as defined in claim 1 wherein at least one of said transmit elements is an antenna.

8. The invention as defined in claim 1 wherein L=4 and said time sequences are spread and arranged according to a matrix, each spread time sequence being a row of said matrix and being transmitted by a respective one of said transmit elements, said matrix being arranged as follows:

$$\begin{bmatrix} b_1\overline{c}_1 & b_2^*\overline{c}_2 & b_3\overline{c}_3 & b_4^*\overline{c}_4 \\ b_2\overline{c}_1 & -b_1^*\overline{c}_2 & -b_4\overline{c}_3 & b_3^*\overline{c}_4 \\ b_3\overline{c}_1 & b_4^*\overline{c}_2 & -b_1\overline{c}_3 & -b_2^*\overline{c}_4 \\ b_4\overline{c}_1 & -b_3^*\overline{c}_2 & b_2\overline{c}_3 & -b_1^*\overline{c}_4 \end{bmatrix}$$

where:
b_1, b_2, b_3, and b_4 are said symbol derivatives from data substreams 1, 2, 3, and 4, respectively;
* indicates complex conjugate; and
$\overline{C}_l$, l=1 .., L are each horizontal vectors of a spreading code, each of said horizontal vectors spans 1 symbol period and contains N chips, where N is the spreading gain.

9. A transmitter adapted for use with four transmit elements to transmit a source bit stream, comprising:

means for dividing said source bit stream into L data substreams, L > 2;
means grouping derivatives of symbols derived from each of said data substreams to form four transmit time sequences, each of said time sequences spanning L symbol periods, at least one of said derivatives of said symbols being a complex conjugate of one of said symbols; and
means for grouping said time sequences into a matrix, each time sequence being a row of said matrix.

10. The invention as defined in claim 9 further comprising L means for encoding each of said data substreams prior to symbols of said data substreams being grouped by said means for grouping, so that said encoded data substreams are grouped by said means for grouping.

FIG. 1

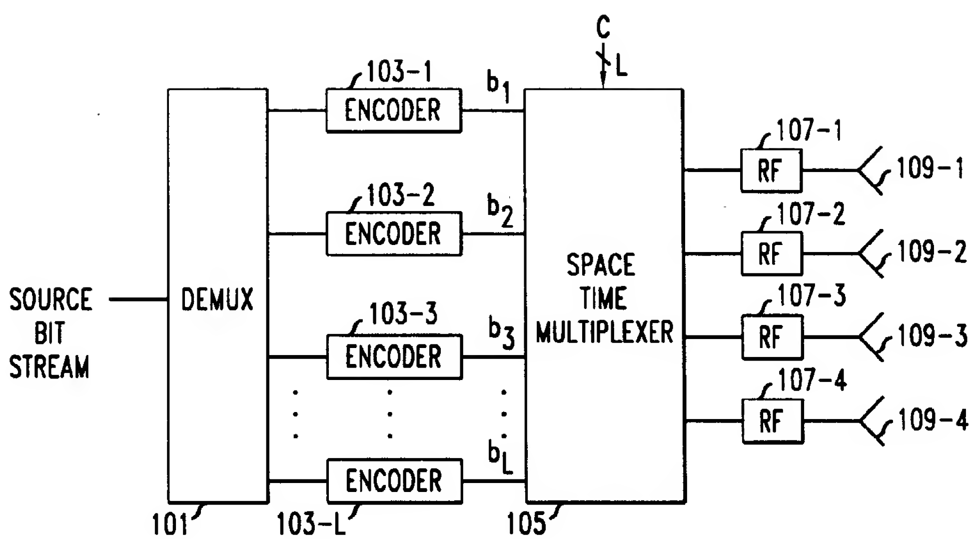
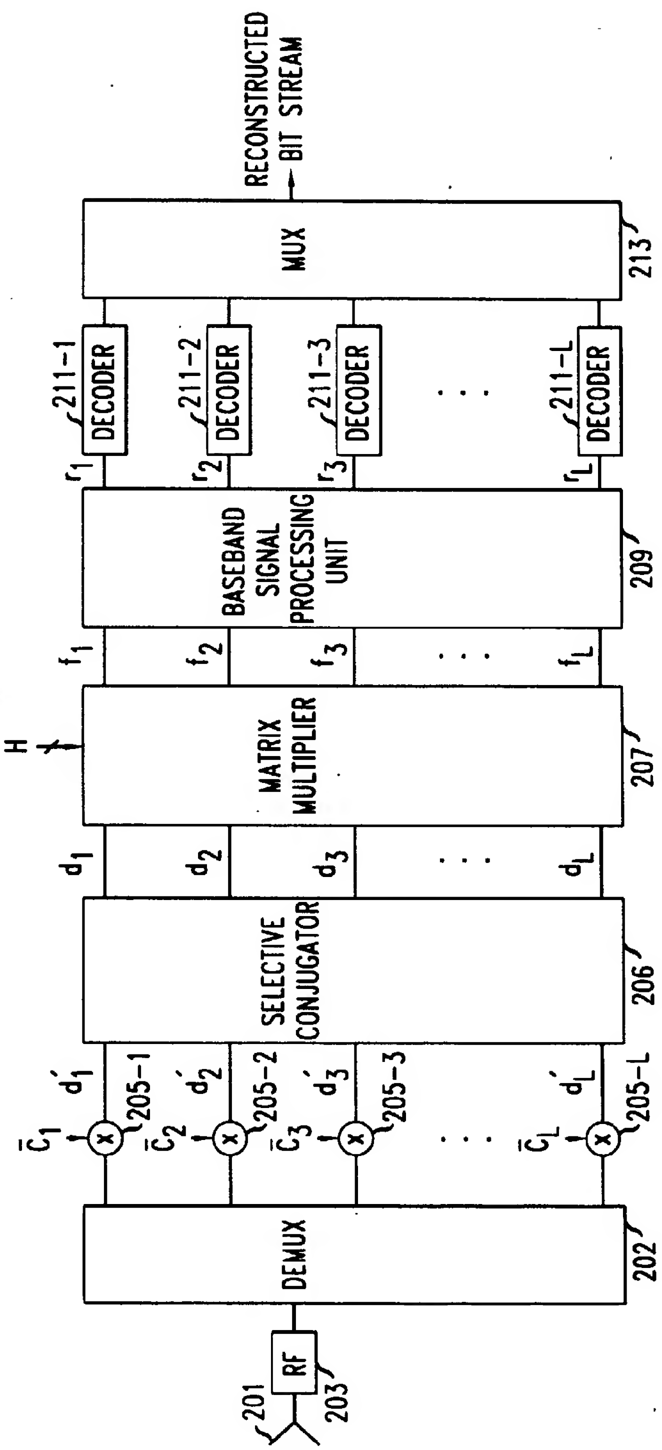

C
L
103-1
ENCODER
b1
107-1
RF
109-1
103-2
ENCODER
b2
107-2
RF
109-2
SPACE
TIME
MULTIPLEXER
SOURCE
BIT
STREAM
DEMUX
103-3
ENCODER
b3
107-3
RF
109-3
107-4
RF
109-4
bL
ENCODER
101
103-L
105

FIG. 2


RECONSTRUCTED BIT STREAM
MUX
213
DECODER
211-1
211-2
211-3
211-L
BASEBAND SIGNAL PROCESSING UNIT
209
MATRIX MULTIPLIER
207
H
SELECTIVE CONJUGATOR
206
205-1
205-2
205-3
205-L
DEMUX
202
RF
201
203

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 30 7260

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X	WO 00 14921 A (AT & T CORP) 16 March 2000 (2000-03-16) * page 17, line 15 - line 26 * * figure 2 *	1,3-7,9, 10	H04L1/06
X	EP 0 905 920 A (LUCENT TECHNOLOGIES INC) 31 March 1999 (1999-03-31) * page 6, line 54 - page 7, line 49 * * page 8, line 56 - line 57 * * page 9, line 16 - line 18 * * figure 4 *	1,3-7,9, 10	
X	TAROKH ET AL: "Space-time block codes from orthogonal designs" IEEE TRANSACTIONS ON INFORMATION THEORY , vol. 45, no. 5, July 1999 (1999-07), pages 1456-1467, XP002161393 IEEE, NEW YORK, US ISSN: 0018-9448 * section V-F *	1,3-7,9, 10	
			TECHNICAL FIELDS SEARCHED (Int.Cl.7)
P,X	PAPADIAS C B ET AL: "A space-time coding approach for systems employing four transmit antennas" 2001 IEEE INTERNATIONAL CONFERENCE ON ACOUSTICS, SPEECH, AND SIGNAL PROCESSING. PROCEEDINGS, SALT LAKE CITY, UT, USA, vol. 4, 7 - 11 May 2001, pages 2481-2484, XP002195791 2001, USA, IEEE ISBN: 0-7803-7041-4 * the whole document *	1-10	H04L

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	11 April 2002	Ghigliotti, L

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 01 30 7260

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report.
The members are as contained in the European Patent Office EDP file on
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11-04-2002

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			EP	1110344	A1	27-06-2001
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			JP	11168453	A	22-06-1999

EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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(54) Multiplexing method in a multicarrier transmit diversity system

(57) The invention relates to a method of multiplexing data words in a multicarrier transmit diversity system. The method comprises the step of generating a plurality of data blocks, each data block comprising data words and each data word containing data symbols derived from a data signal, the step of determining for one or more data blocks in dependence on at least one transmission constraint if the data words of said one or more data blocks are to be multiplexed in the time domain or in the frequency domain and the step of multiplexing the data words of the data blocks in accordance with the determination result.

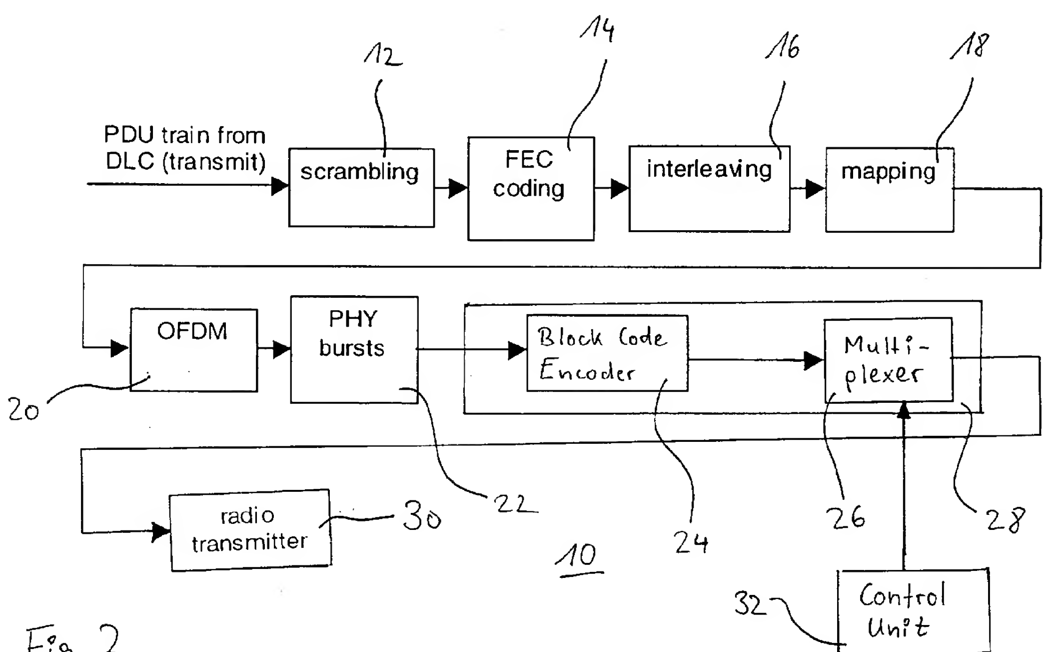

Fig. 2

EP 1 241 824 A1

Description

BACKGROUND OF THE INVENTION

Technical Field

[0001] The present invention relates to the field of transmit antenna diversity and in particular to a method of multiplexing data words in a multi carrier transmit diversity system. The invention also relates to a multiplexer for multiplexing a sequence of data symbols and a demultiplexer for demultiplexing a multiplexed sequence of data symbols.

Discussion of the Prior Art

[0002] Peak transmission rates in wireless communication systems have steadily increased during the last years. However, peak transmission rates are still limited for example due to path loss, limited spectrum availability and fading.

[0003] Transmitter diversity is a highly effective technique for combating fading in wireless communications systems. Several different transmit diversity schemes have been proposed. In Li, Y.; Chuang, J.C.; Sollenberger, N.R.: Transmitter diversity for OFDM systems and its impact on high-rate data wireless networks, IEEE Journal on Selec. Areas, Vol. 17, No. 7, July 1999 the transmit diversity schemes of delay, permutation and space-time coding are exemplarily described. According to the delay approach, a signal is transmitted from a first transmitter antenna and signals transmitted from further transmitter antennas are delayed versions of the signal from the first transmitter antenna. In the permutation scheme, the modulated signal is transmitted from a first transmitter antenna and permutations of the modulated signal are transmitted from further transmitter antennas. By means of space-time coding a signal is encoded into several data words and each data word is transmitted from a different transmitter antenna. During transmission the data words are spread (or multiplexed) in the time domain by successively transmitting the data symbols of a data word over a single carrier frequency.

[0004] A further transmit diversity scheme for a multicarrier system is space-frequency coding. By means of space-frequency coding a signal is encoded into several data words and each data word is spread (or multiplexed) in the frequency domain by transmitting the data symbols of each data word on orthogonal frequencies, i.e. orthogonal subcarriers. An exemplary scheme for space-frequency coding is described in Mudulodu, S.; Paulraj, A.: A transmit diversity scheme for frequency selective fading channels, Proc. Globecom, San Francisco, pp. 1089-1093, Nov. 2000. According to the multicarrier system described in this paper, the data words relating to an encoded signal are preferably multiplexed in the time domain although orthogonal frequencies are available and the data words could thus also be multiplexed in the frequency domain. This is due to the fact that if multiplexing in the frequency domain is utilized the employed frequencies, i.e. subcarriers, must see the same channel, which may not always be possible in a frequency selective fading channel. However, in case the subcarriers experience the same channel, it is stated that either multiplexing in the time domain or multiplexing in the frequency domain or a combination of the two may be used. By combining multiplexing in the time domain and in the frequency domain the data symbols of a data word are simultaneously multiplexed in the time domain and in the frequency domain. This means that the data word is spread both across time and across frequencies.

[0005] Departing from the various transmit diversity schemes hitherto known there is a need for a method of multiplexing data words in a multicarrier transmit diversity system which can easily be adapted to the specifications of different wireless communications systems. There is also a need for a corresponding multiplexer and a demultiplexer.

BRIEF DESCRIPTION OF THE INVENTION

[0006] The existing need is satisfied by a method of multiplexing data words in a multicarrier transmit diversity system which comprises the step of generating a plurality of data blocks, each data block comprising data words and each data word containing data symbols derived from a data signal, the step of determining for one or more of the data blocks in dependence on at least one transmission constraint if the data words of said one of more data blocks are to be multiplexed in a time domain or in a frequency domain and the step of multiplexing the data words of the data blocks in accordance with the result of the determination.

[0007] The multiplexing method of the invention is not restricted to a specific transmit diversity scheme as long as the utilized transit diversity scheme enables to generate from a data signal a plurality of data blocks having the above structure. For example, the transmit diversity schemes of block coding and of permutation allow to generate such data blocks. Preferably, the generated data blocks have the structure of a matrix similar to a space-time block code (STBC) matrix. Also, it is not required that the transmit diversity scheme guarantees full transmit diversity. In other words: the invention does not necessitate that each information symbol comprised within the data signal is transmitted from each transmitter antenna. Nonetheless, a preferred embodiment of the invention comprises the feature of full transmit di-

versity.

[0008] Moreover, the invention is not restricted to any number of transmit and receive antennas. Preferably, the number of data words per data block equals the number of transmit antennas such that each data word of a data block may be transmitted from an individual transmitter antenna. If more than one receive antenna is provided, the receive diversity scheme of maximum-ratio combining can be applied. However, other receive diversity schemes may be used as well.

[0009] According to the invention, it is decided on a data block level how the data words are to be multiplexed. The decision on the data block level allows to change the multiplexing domain from one data block to a subsequent data block which is advantageous if one has to cope with specific predefined or varying transmission constraints. Also, the multiplexing method according to the invention can be applied in various wireless communication systems without major changes due to the specific multiplexing flexibility gained by selecting the multiplexing domain on the data block level. The multiplexing domain can be determined for each data block individually or simultaneously for a plurality of data blocks. For example, it can be decided for a sequence of data blocks that all data words comprised within the sequence of data blocks are to be multiplexed in either the time domain or in the frequency domain.

[0010] The multiplexing domain is determined by taking into account one or more transmission constraints. For example, the transmission constraints may comprise one or more physical transmission constraints or one or more data-related transmission constraints. It can also comprise both one or more physical transmission constraints and one or more data-related transmission constraints. The physical transmission constraints relate to the physical transmission conditions and can be derived from physical transmission parameters like a channel coherence bandwidth or a coherence time. The data-related transmission constraints relate to system specific constraints regarding for example the employed multiplexing scheme for the data words, the structure of the data signal, the structure of the data blocks, the structure of the data words or the structure of the data symbols.

[0011] The data symbols may be derived from the data signal in various ways dependent on the transmit diversity scheme which is used. If, for example, the transmit diversity scheme of permutation is used, the data symbols contained in the data words are permutations of information symbols comprised within the data signal. As a further example, if the transmit diversity scheme of block coding is used, the data symbols contained in the data words are obtained from the information symbols comprised within the data signal by means of permutation and basic arithmetic operations, such as negation and complex conjugation.

[0012] The data signal from which the one or more data blocks are generated can have any format. According to a preferred embodiment, the data signal has the format of a sequence of discrete information symbols. For example, the data signal may have the structure of vectors, each vector comprising a predefined number of information symbols. The nature of the information symbols may depend on the specific wireless communication system in which the multiplexing method according to the invention is used. Many wireless communication systems employ different types of information symbols for different purposes. For example, some wireless communication systems use data signals which comprise a preamble, one or more user data sections or both a preamble and one or more user data sections. Usually, the preamble has a predefined structure and is utilized for purposes like channel estimation, frequency synchronization and timing synchronization.

[0013] In the following, several exemplary data-related transmission constraints are described in more detail. According to a first embodiment, the data-related transmission constraint is a predefined number N of data symbols to be comprised within each data word which is to be multiplexed in the time domain. Usually, the number N of data symbols to be comprised within each data word cannot arbitrarily be chosen because it may depend for example on a code rate, on the condition that the data blocks have to be orthogonal matrices or on the availability of memory resources within the multicarrier transmit diversity system.

[0014] When the data words of a specific data block are to be multiplexed in the time domain, the number N of data symbols to be comprised within each data word may represent the number of time slots required for the transmission of a single data word over a single subcarrier. On the other hand, when the data words of a specific data block are to be multiplexed in the frequency domain, the number N of data symbols to be comprised within each data word stands for the number of subcarriers required to transmit a single data word during a single time slot.

[0015] Preferably, all data words of an individual data block comprise the same number of data symbols. If the data signal has such a structure that the number of data symbols comprised within each data word of a specific data block equals the predefined number N of data symbols, the data words of this data block may be multiplexed in the time domain. Otherwise, i.e. if the data signal has such a structure that the number of data symbols comprised within each data word of a specific data block does not equal the predefined number N of data symbols, the data words of this data block may be multiplexed in the frequency domain. Such a distinction will become necessary if the data signal or a portion thereof has a predefined length because the predefined length may imply that the total number N_D of data symbols which corresponds to the predefined length of the data signal or a portion thereof is not an integer multiple of the predefined number N of data symbols which should be comprised within a data word to be multiplexed in the time domain. In such a situation integer multiples of the predefined number N of data symbols are arranged in data blocks

of data words which are multiplexed in a time domain and a remainder $N_R = \mathrm{mod}(N_D/N)$ of data symbols is arranged in a data block with data words which are multiplexed in the frequency domain.

[0016] Thus, by combining multiplexing in the time domain and in the frequency domain, data symbol fitting problems resulting from the predefined number N of data symbols to be comprised within each data word which is to be multiplexed in the time domain can be solved. Such data symbol fitting problems may for example become relevant when the data signal or a portion of the data signal has a predefined length because the wireless transmission system necessitates that the preamble portion or the user data portion of a data signal comprises a certain number of information symbols. Thus the data words of all data blocks except for the last data block are multiplexed in the time domain and the data words of the last data block are either multiplexed in the time domain or in the frequency domain depending on whether or not the data words of the last data block contain a number of data symbols which equals the predefined number N of data symbols.

[0017] So far the data-related transmission constraint of a predefined number N of data symbols to be comprised within each data word has been illustrated. According to a second embodiment, the data signal may comprise one or more periodic structures and the data related transmission constraint may be a preservation of the periodic structures such that the periodic structures are still periodic on a receiver side. The one or more periodic structures may be comprised within a preamble of the data signal, for example in the form of two or more identical preamble information symbols. Periodic structures are advantageous because they allow the use of synchronization algorithms with comparatively low complexity.

[0018] In case of multiplexing data symbols relating to periodic structures in the time domain the periodicity of the periodic structures may get lost. Therefore, at least the data words of data blocks which relate to the periodic structures or parts of periodic structures are multiplexed in the frequency domain. By multiplexing the data words of these data blocks in the frequency domain it can be ensured that the periodicity of the periodic structures is maintained.

[0019] When the data words of data blocks generated from periodic structures or portions thereof are multiplexed in the frequency domain, the data words of data blocks generated from the remaining data signal are preferably multiplexed in the time domain. If, for example, the data words of data blocks generated from a preamble comprising periodic structures are multiplexed in the frequency domain, the data words of data blocks generated from a corresponding user data section may be multiplexed at least partly in the time domain.

[0020] Instead of data-related transmission constraints or in addition to data-related transmission constraints physical transmission constraints can be taken into account when deciding if the data words of one or more specific data blocks are to be multiplexed in the time domain or in the frequency domain. According to a preferred embodiment, the decision is made based on simultaneously evaluating a combination of one or more data-related transmission constraints and one or more physical transmission constraints.

[0021] The physical transmission constraints may be determined based on at least one of a channel coherence bandwidth

$$B_C \approx 1/\tau_{rms} \tag{1}$$

and a coherence time

$$t_C \approx 1/(2 \cdot f_D) \tag{2}$$

wherein f_D is the doppler frequency and τ_{rms} is the root mean square of the delay spread of the channel impulse response.

[0022] Many transmit diversity schemes require constant or at least approximately constant channel parameters during transmission of one data word. If the data words are to be multiplexed in the frequency domain, a comparatively large coherence band width is required. This means that the relation

$$B_C \gg N/T \tag{3}$$

has to be fulfilled at least approximately, wherein N is the number of data symbols per data word and T is the duration of one of the data symbols, i.e. the duration of one time slot. A comparatively large coherence bandwidth requires that the channel parameters of N adjacent subcarriers have to be almost constant.

[0023] On the other hand, if the data words are to be multiplexed in the time domain, a comparatively large coherence time is required. This means that the relation

$$t_C \gg T \cdot N \quad (4)$$

has to be fulfilled at least approximately. In other words: N subsequent data symbols have to have nearly constant channel parameters, i.e. the channel parameters for a single subcarrier have to remain constant for a period of $N \cdot T$.

[0024] The physical transmission constraint may be determined by assessing if one or both of the relations (3) and (4) are fulfilled. Dependent on which of the two relations (3) and (4) is fulfilled best it is decided that the data words of the data blocks are to be multiplexed either in the time domain or in the frequency domain as a general rule. Deviations from this general rule may become necessary due to data-related transmission constraints. For example, the data symbol fitting problem or the problem encountered with periodic structures may necessitate that although multiplexing in the time domain is generally to be preferred, the data words of at least some data blocks have to be multiplexed in the frequency domain. As a further example, changing transmission conditions may necessitate that the data words of some data blocks have to be multiplexed in the time domain and the data words of other data blocks have to be multiplexed in the frequency domain. As a third example, the data words of data blocks generated from a preamble may be multiplexed in the time domain and the data words of data blocks generated from a user data section may be multiplexed in the frequency domain. Such a combination has the advantage that the above-mentioned data symbol fitting problem, which usually is most relevant for the user data section, can be avoided while the multiplexing in the time domain of the data words of data blocks generated from the preamble allows a good channel estimation.

[0025] It was mentioned above that in order to achieve full diversity each information symbol has to be transmitted from each transmitter antenna. A further requirement of full transmit diversity is that the antenna signals are orthogonal to each other. This means that the data symbols have to be modulated onto subcarriers which are orthogonal to each other. However, the invention can also be practized in case the subcarriers are not orthogonal.

BRIEF DESCRIPTION OF THE DRAWINGS

[0026] Further advantages of the invention will become apparent by reference to the following description of a preferred embodiment of the invention in the light of the accompanying drawings, in which :

- Fig. 1 shows a data signal in the form of a physical burst to be processed in accordance with the invention;
- Fig. 2 is a block diagram of a transceiver for wireless communication adapted to multiplex data words in accordance with the invention;
- Fig. 3 shows several modulation schemes defined in the HIPERLAN/2 standard;
- Fig. 4 shows the block code encoder of the transceiver depicted in Fig. 2;
- Fig. 5 shows the configuration of a transmit antenna diversity scheme;
- Fig. 6 is a schematic diagram of multiplexing data words in the time domain in accordance with the invention; and
- Fig. 7 is a schematic diagram of multiplexing data words in the frequency domain in accordance with the invention.

DESCRIPTION OF PREFERRED EMBODIMENTS

[0027] Although the present invention can be used in any multicarrier transmit diversity system which employs a transmit diversity scheme allowing to generate data blocks having a structure as described above, the following description of preferred embodiments is exemplarily set forth with respect to a multicarrier system which employs orthogonal frequency division multiplexing (OFDM) and which utilizes block coding for generating data blocks from a data signal.

[0028] The exemplary multicarrier system described below is derived from the European wireless local area network (WLAN) standard high performance radio local area network type 2 (HIPERLAN/2). HIPERLAN/2 systems are intended to be operated in the 5 GHz frequency band. A system overview of HIPERLAN/2 is given in ETSI TR 101 683, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; System Overview, V1.1.1 (2000-02) and the physical layer of HIPERLAN/2 is described in ETSI TS 101 475; Broadband Radio Access networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, V1.1.1 (2000-04). The multicarrier scheme of OFDM, which is specified in the HIPERLAN/2 standard, is very robust in frequency selective environments.

[0029] Up to now, the HIPERLAN/2 system and many other wireless communications systems do not support transmit diversity in spite of the fact that transmit diversity would improve the transmission performance and reduce negative effects of fast fading like Rayleigh fading. However, applying standard transmit diversity schemes to multicarrier communications systems may lead to various problems which are hereinafter exemplarily described with respect to the HIPERLAN/2 system.

[0030] In Fig. 1 a typical physical burst of HIPERLAN/2 is illustrated. The physical burst comprises a preamble consisting of preamble symbols and a user data section consisting of user data symbols. In HIPERLAN/2 five different physical bursts are specified and each kind of physical burst has a unique preamble. However, the last three preamble symbols constitute a periodic structure which is identical for all preamble types. This periodic structure consists of a short OFDM symbol C32 of 32 samples followed by two identical regular OFDM symbols C64 of 64 samples. The short OFDM symbol C32 is a cyclic prefix which is a repetition of the second half of one of the C64 OFDM symbols. The so-called C-preamble depicted in Fig. 1 is used in HIPERLAN/2 for channel estimation, frequency synchronization and timing synchronization. The periodic structure within the C-preamble is necessary in order to allow the use of synchronization algorithms with comparatively low complexity.

[0031] The user data section of the physical burst depicted in Fig. 1 comprises a variable number N_{SYM} of OFDM symbols required to transmit a specific protocol data unit (PDU) train. Each OFDM symbol of the user data section consists of a cyclic prefix and a useful data part. The cyclic prefix consists of a cyclic continuation of the useful data part and is inserted before it. Thus, the cyclic prefix is a copy of the last samples of the useful data part. The length of the useful data part is equal to 64 samples and has a duration of 3,2 μs. The cyclic prefix has a length of either 16 (mandatory) or 8 (optional) samples and a duration of 0,8 μs or 0,4 μs, respectively. Altogether, a OFDM symbols thus has a length of either 80 or 72 samples corresponding to a symbol duration of 4,0 μs or 3,6 μs, respectively. An OFDM symbol therefore has an extension in the time domain. A OFDM symbol further has an extension in the frequency domain. According to HIPERLAN/2, a OFDM symbol extends over 52 subcarriers. 48 subcarriers are reserved for complex valued subcarrier modulation symbols and 4 subcarriers are reserved for pilots.

[0032] From the above it becomes clear that the HIPERLAN/2 physical burst depicted in Fig. 1 has a predefined length both in a time direction and in a frequency direction. Moreover, the physical burst of Fig. 1 comprises a periodic structure. It are among others these features of the physical burst of Fig. 1 which may lead to problems when the HIPERLAN/2 system or a similar wireless communication system has to be adapted to transmit diversity.

[0033] For typical HIPERLAN/2 scenarios the above relation (4) is usually fulfilled because the doppler frequency f_D is comparatively low. However, especially in outdoor environments, relatively large delay spreads can occur. Consequently, relation (3) cannot always be fulfilled. Therefore, a transmit diversity scheme like STBC multiplexing in the time domain should generally be a preferred transmit diversity scheme for a HIPERLAN/2 scenario from the point of view that the channel over one space-time data word should be as constant as possible. However, severe problems arise when STBC is applied to physical bursts having the structure depicted in Fig. 1 or a similar structure.

[0034] Both the physical burst and the OFDM symbols comprised therein have predefined dimensions in the time domain and in the frequency domain. Concurrently, STBC requires that each STBC data word has a predetermined length N. Thus, data unit fitting problems arise if the dimension of e.g. an OFDM symbol of the preamble or of the user data section cannot be mapped on an integer multiple of the length of one STBC data word. Moreover, when applying STBC to the periodic C-preamble depicted in Fig. 1, the periodicity of the C-preamble gets lost. This is due to the fact that the one or more STBC data words relating to the second C64 OFDM symbol will no longer be equal to the one or more STBC data words relating to the first C64 OFDM symbol. The loss of periodicity, however, leads to the problem that the symbol synchronization algorithms which make use of a periodic structure within the preamble can no longer be employed. Also, the C32 OFDM symbol cannot serve any longer as a guard interval separating the OFDM symbols within the preamble. The reason therefore is that in case of multipath propagation the first C64 OFDM symbol interferes with the second C64 OFDM symbol which is no longer equal to the first C64 OFDM symbol.

[0035] The above problems and further problems not explicitly discussed above do not occur when the data words are multiplexed in accordance with the invention. In Fig. 2, the physical layer of a transceiver 10 which is adapted to implement the method according to the invention is illustrated. The transceiver 10 comprises a scrambler 12, an FEC coding unit 14, an interleaving unit 16, a mapping unit 18, an OFDM unit 20, a burst forming unit 22, a block code encoder 24, a multiplexer 26, a radio transmitter 30 and a control unit 32. The block code encoder 24 and the multiplexer 26 together form an encoder/multiplexer unit 28.

[0036] The transceiver 10 depicted in Fig. 1 receives as input signal a PDU train from a data link control (DLC). Each PDU train consists of information bits which are to be framed into a physical burst, i.e. a sequence of OFDM symbols to be encoded, multiplexed and transmitted.

[0037] Upon receipt of a PDU train the transmission bit rate within the transceiver 10 is configured by choosing an appropriate physical mode based on a link adaption mechanism. A physical mode is characterized by a specific modulation scheme and a specific code rate. In the HIPERLAN/2 standard several different coherent modulation schemes like BPSK, QPSK, 16-QAM and optional 64-QAM are specified. Also, for forward error control, convolutional codes

with code rates of 1/2, 9/16 and 3/4 are specified which are obtained by puncturing of a convolutional mother code of rate 1/2. The possible resulting physical modes are depicted in Fig. 3. The data rate ranging from 6 to 54 Mbit/s can be varied by using various signal alphabets for modulating the OFDM subcarriers and by applying different puncturing patterns to a mother convolutional code.

[0038] Once an appropriate physical mode has been chosen, the N_{BPDU} information bits contained within the PDU train are scrambled with the length-127 scrambler 12. The scrambled bits are then output to the FEC coding unit 14 which encodes the N_{BPDU} scrambled PDU bits according to the previously set forward error correction.

[0039] The encoded bits output by the FEC coding unit 14 are input into the interleaving unit 16 which interleaves the encoded bits by using the appropriate interleaving scheme for the selected physical mode. The interleaved bits are input into the mapping unit 18 where sub-carrier modulation by mapping the interleaved bits into modulation constellation points in accordance with the chosen physical mode is performed. As mentioned above, the OFDM subcarriers are modulated by using BPSK, QPSK, 16-QAM or 64-QAM modulation depending on the physical mode selected for data transmission.

[0040] The mapping unit 18 outputs a stream of complex valued subcarrier modulation symbols which are divided in the OFDM unit in groups of 48 complex numbers. In the OFDM unit a complex base band signal is produced by OFDM modulation as described in ETSI TS 101 475, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, V1.1.1 (2000-04).

[0041] The complex base band OFDM symbols generated within the OFDM unit 20, where pilot subcarriers are inserted, are input into the physical burst unit 22, where an appropriate preamble is appended to the PDU train and the physical burst is built. The physical burst produced by the physical burst unit 22 has a format as depicted in Fig. 1. The physical burst unit 22 thus outputs a sequence of complex base band OFDM symbols in the form of the physical burst to the block code encoder 24.

[0042] The function of the block code encoder 24 is now generally described with reference to Fig. 4. In general, the block code encoder 24 receives an input signal in the form of a sequence of vectors $\mathbf{X} = [X_1 X_2 ... X_K]^T$ of the length K. The block code encoder 24 encodes each vector $\mathbf{X}$ and outputs for each vector $\mathbf{X}$ a data block comprising a plurality of signal vectors $\mathbf{C}^{(1)}, \mathbf{C}^{(2)} ..., \mathbf{C}^{(M)}$ as depicted in Fig. 4. Each signal vector $\mathbf{C}^{(1)}, \mathbf{C}^{(2)} ..., \mathbf{C}^{(M)}$ corresponds to a single data word. Thus, the data block generated from the vector $\mathbf{X}$ comprises M data words wherein M is the number of transmitter antennas.

[0043] Each data word $\mathbf{C}^{(i)}$ with i = 1...M comprises N data symbols, i.e. each data word $\mathbf{C}^{(i)}$ has a length of N. The value of N cannot be freely chosen since the matrix $\mathbf{C}$ spanned by the data words $\mathbf{c}^{(i)}$ has to be orthogonal in this embodiment. Several examples for data blocks in the form of orthogonal code matrices $\mathbf{C}$ are described in US 6,088,408. In the block coding approach described in the present embodiment all data symbols c_j^i of the code matrix $\mathbf{C}$ are derived from the components of the input vector $\mathbf{X}$ and are simple linear functions thereof or of its complex conjugate.

[0044] If a receive signal vector $\mathbf{Y}$ at one receive antenna is denoted by $\mathbf{Y} = [Y_1 Y_2 ... Y_N]^T$, the relationship between $\mathbf{Y}$ and the code matrix $\mathbf{C}$ is as follows:

$$\begin{bmatrix} Y_1 \\ Y_2 \\ \cdots \\ Y_N \end{bmatrix} = \begin{bmatrix} c_1^{(1)} & c_1^{(2)} & \cdots & c_1^{(M)} \\ c_2^{(1)} & \cdots & & c_2^{(M)} \\ \cdots & & \cdots & \cdots \\ c_N^{(1)} & c_N^{(2)} & \cdots & c_N^{(M)} \end{bmatrix} \cdot \begin{bmatrix} h^{(1)} \\ h^{(2)} \\ \cdots \\ h^{(M)} \end{bmatrix} \qquad (5)$$

where $h^{(i)}$ represents the channel coefficient of the channel from the i-th transmit antenna to the receive antenna. A generalization to more receive antennas is straightforward.

[0045] In the following examples of possible block code matrices for two and three transmitter antennas, respectively, are discussed in more detail. The configuration of a wireless communication system with two transmit antennas and one receive antenna is depicted in Fig. 5. For two transmit antennas one possible block code matrix $\mathbf{C}$ with a code rate R = 1 is :

$$\mathbf{C} = \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_2^* \end{bmatrix} \qquad (6)$$

[0046] For three transmit antennas one possible block code matrix **C** with a code rate R = 0,5 is:

$$\mathbf{C} = \begin{bmatrix} X_1 & X_2 & X_3 \\ -X_2 & X_1 & -X_4 \\ -X_3 & X_4 & X_1 \\ -X_4 & -X_3 & X_2 \\ X_1^* & X_2^* & X_3^* \\ -X_2^* & X_1^* & -X_4^* \\ -X_3^* & X_4^* & X_1^* \\ -X_4^* & -X_3^* & X_2^* \end{bmatrix} \qquad (7)$$

[0047] The code rate R is defined as the ratio of the length K of the input vector **X** and the length N of each code word $\mathbf{C}^{(i)}$:

$$R = K/N \qquad (8)$$

[0048] As can be seen from Fig. 4, the block code encoder 24 outputs for each data signal in the form of a vector **X** a data block in the form of a matrix **C**. The data block output by the block code encoder 24 is input into the multiplexer 26 which multiplexes the data words (vectors $\mathbf{C}^{(i)}$) of each data block in accordance with an externally provided control signal either in the time domain or in the frequency domain. The control signal is generated by the control unit 32 based on an assessment of the transmission constraints. The assessment of the transmission constraints and the controlling of the multiplexer 26 by means of the control unit 32 will be described later in more detail.

[0049] In the multicarrier scheme OFDM, the output of the block code encoder 24 is modulated onto subcarriers which are orthogonal to each other. There exist essentially two possibilities to multiplex a data block comprising individual data words in an OFDM system. According to a first possibility depicted in Fig. 6, the data words of a specific data block are extended in the time direction (STBC). In other words: The data words are multiplexed in the time domain. According to a second possibility, the data words of a data block are extended in the frequency direction as depicted in Fig. 7. This means that the data words are multiplexed in the frequency domain. Multiplexing the data words of a data block in the form of a code matrix in the frequency domain will in the following be referred to as space-frequency block coding (SFBC).

[0050] As can be seen from Figs. 6 and 7, the individual data words of a data block are transmitted from different transmit antennas. According to the multiplexing scheme of Fig. 6, an individual data block is transmitted on an individual subcarrier over a time interval of $N \cdot T$, wherein N is the number of data symbols per data word and T is the duration of one of the data symbols. According to the multiplexing scheme of Fig. 7, an individual data block is spread over N subcarriers and is transmitted during a time interval of T. It can clearly be seen that the multiplexing scheme of Fig. 6 can generally be employed when the relation (4) is fulfilled and the multiplexing scheme of Fig. 7 can generally be employed when the relation (3) is satisfied.

[0051] The encoded and multiplexed output signal of the encoder/multiplexer unit 28 is input into the radio transmitter 30. The radio transmitter 30 performs radio transmission over a plurality of transmit antennas by modulating a radio frequency carrier with the output signal of the encoder/multiplexer unit 28. The transceiver 10 of Fig. 2 further comprises a receiver stage not depicted in Fig. 2. The receiver stage has a physical layer with components for performing the inverse operations of the components depicted in Fig. 2. For example, the receiver stage comprises a descrambler, a FEC decoding unit, a demultiplexer/decoder unit with a demultiplexer and a block code decoder, etc.

[0052] Now, the control of the multiplexer 26 will be described in more detail with reference to both physical and data-related transmission constraints that may occur if physical bursts as the one depicted in Fig. 1 are employed. In accordance with typical HIPERLAN/2 scenarios, it is supposed that relation (4) is fulfilled and that it cannot always be guaranteed that relation (3) is fulfilled. This corresponds to the realistic situation that the basic performance of STBC transmission is better than the basic performance of SFBC transmission. Basic performance here means that only physical transmission constraints are taken into account. In such a case the control unit 32 may decide that the data blocks have to be multiplexed in the time domain. However, if the physical transmission parameters change, there might occur the case where relation (4) is no longer fulfilled whereas relation (3) is fulfilled at least approximately. In

this case the control unit 32 will decide that the data words of the data blocks are no longer multiplexed in the time domain. Instead, the control unit 32 controls the multiplexer 26 such that the data words of the data blocks are multiplexed in the frequency domain.

[0053] So far only physical transmission constraints have been considered. Should data-related transmission constraints also be of importance, the control unit 32 controls the multiplexer 26 by additionally taking into account data-related transmission constraints.

[0054] It has been mentioned above that the transmission constraints which have to be considered in context with the physical burst depicted in Fig. 1 are the preservation of a periodic structure in the C-preamble and the provision of a predefined number N of data symbols in each data word which is to be multiplexed in the time domain. These two data-related transmission constraints can occur in several combinations.

[0055] According to a first scenario, the data signal has the structure of the physical burst depicted in Fig. 1 and comprises a user data section and a preamble with a periodic structure. It is further supposed that the data-related transmission constraint of preserving the periodic structure has to be taken into account while no data symbol fitting problem occurs with respect to the user data section. In such a case the data words of data blocks relating to the preamble are multiplexed in accordance with SFBC in the frequency domain and the data words of data blocks relating to the user data section are multiplexed in accordance with STBC in the time domain. By multiplexing the data words derived from the preamble in the frequency domain a preservation of the order of the C32 OFDM symbols and the two C64 OFDM symbols can be achieved.

[0056] According to a second scenario derived from the physical burst depicted in Fig. 1, the periodic structure within the preamble has to be preserved and additionally the data symbol fitting problem has to be taken into account with respect to the user data section. Like in the first scenario, the data words of data blocks derived from the preamble are multiplexed in accordance with SFBC in the frequency domain. Due to the data symbol fitting problem the data words of the last data block relating to the user data structure contains less than the predefined number N of data symbols contained in data words of the previous data blocks. Therefore, only the data words (containing the predefined number N of data symbols) of the previous data blocks are multiplexed in accordance with STBC in the time domain. The data words of the last data block contain $N_R = \mathrm{mod}(N_D/N)$ data symbols and are multiplexed in accordance with SFBC in the frequency domain, wherein N_D is the total number of data symbols to be transmitted over one transmit antenna.

[0057] According to a third scenario, the data-related transmission constraint of the preservation of a periodic structure within the preamble is not relevant but the data symbol fitting problem is relevant with respect to the user data section. In this case the data words of data blocks relating to the preamble are multiplexed in accordance with STBC in the time domain and the data words of data blocks relating to the user data section are multiplexed as described above for the second scenario. In other words: The data words of the last data block have a length of N_R data symbols and the data words of the previous data blocks have the predefined length of N data symbols.

[0058] According to a fourth scenario, the data-related transmission constraint of preserving a periodic structure has not to be taken into account and the physical transmission constraint of $B_C \gg N/T$ is at least approximately fulfilled. In this case the data words of data blocks relating to the preamble are multiplexed in accordance with STBC in the time domain and the data words of data blocks relating to the user data section are multiplexed in accordance with SFBC in the frequency domain. By using STBC for the preamble a good channel estimation can be performed. Due to the use of STBC for the preamble the slightly worse performance of SFBC can be compensated by means of receiver algorithms for interference suppression based on the good channel estimation. Using STBC for the preamble and SFBC for the user data section has the advantage that data symbol fitting problems with respect to the user data section do not appear.

[0059] Additional scenarios based on further combinations of data-related and physical transmission constraints can easily be realized in accordance with the invention. Also, the invention can easily be applied to data signals having a structure different from the structure of the physical burst depicted in Fig. 1. Although the invention is preferably practiced with the transmit diversity scheme of a combination of STBC and SFBC, other transmit diversity schemes can be used as well.

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Claims

1. A method of multiplexing data words in a multicarrier transmit diversity system, comprising:

   a) generating a plurality of data blocks (**C**), each data block (**C**) comprising data words ($\mathbf{C}^{(i)}$) and each data word ($\mathbf{C}^{(i)}$) containing data symbols ($c_j^{(i)}$) derived from a data signal;

   b) determining for one or more data blocks (**C**) in dependence on at least one transmission constraint if the

data words ($\mathbf{C}^{(i)}$) of said one or more data blocks ($\mathbf{C}$) are to be multiplexed in the time domain or in the frequency domain; and

c) multiplexing the data words ($\mathbf{C}^{(i)}$) of the data blocks ($\mathbf{C}$) in accordance with the determination in step b).

2. The method according to claim 1, wherein the data signal comprises at least one of a preamble and a user data section.

3. The method according to claim 1 or 2, wherein the at least one transmission constraint comprises a data-related transmission constraint.

4. The method according to claim 3, wherein the data-related transmission constraint is a predefined number (N) of data symbols ($c_j^{(i)}$) to be comprised within each data word ($\mathbf{C}^{(i)}$) which is to be multiplexed in the time domain.

5. The method according to claim 4, wherein the data words ($\mathbf{C}^{(i)}$) containing the predefined number (N) of data symbols ($c_j^{(i)}$) are multiplexed in the time domain and the data words ($\mathbf{C}^{(i)}$) containing more or less data symbols ($c_j^{(i)}$) are multiplexed in the frequency domain.

6. The method according to claim 4 or 5, wherein the data signal or a portion thereof has a predefined length and wherein integer multiples of the predefined number of data symbols ($c_j^{(i)}$) are arranged in data blocks ($\mathbf{C}$) with data words ($\mathbf{C}^{(i)}$) which are multiplexed in the time domain and a remainder of data symbols ($c_j^{(i)}$) is arranged in data blocks ($\mathbf{C}$) with data words ($\mathbf{C}^{(i)}$) which are multiplexed in the frequency domain.

7. The method according to claim 6, wherein the user data section of the data signal has the predefined length.

8. The method according to claim 7, wherein the data words ($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the preamble are either multiplexed completely in the frequency domain or completely in the time domain dependent on the transmission constraint.

9. The method according to one of claims 1 to 8, wherein the data signal comprises one or more periodic structures (C32, C64).

10. The method according to claim 9, wherein the one or more periodic structures (C32, C64) are contained within the preamble.

11. The method according to claim 9 or 10, wherein the data-related transmission constraint is a preservation of the one or more periodic structures (C32, C64).

12. The method according to one of claims 9 to 11, wherein at least the data words ($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the periodic structures (C32, C64) are multiplexed in the frequency domain.

13. The method according to claim 12, wherein the data words ($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the user data section are multiplexed in the time domain.

14. The method according to one of claims 1 to 13, wherein the at least one transmission constraint comprises a physical transmission constraint.

15. The method according to claim 14, wherein the physical transmission constraint is determined based on at least one of a coherence bandwidth and a coherence time.

16. The method according to claim 15, wherein the physical transmission constraint is determined by assessing if the relationship $B_C >> N/T$ is fulfilled, wherein B_C is the coherence bandwidth, N is the number of data symbols ($c_j^{(i)}$) per data word ($\mathbf{C}^{(i)}$) and T is the

duration of one of the data symbols ($c_j^{(i)}$).

17. The method according to claim 15 or 16,
wherein the physical transmission constraint is determined by assessing if the relationship $t_C >> N \cdot T$ is fulfilled, wherein t_C is the coherence time, N is the number of data symbols ($c_j^{(i)}$) per data word (**C**$^{(i)}$) and T is the duration of one of the data symbols ($c_j^{(i)}$).

18. The method according to claim 16 or 17,
wherein, when the physical transmission constraint $B_c >> N/T$ is at least approximately fulfilled, the data words (**C**$^{(i)}$) of data blocks (**C**) relating to the preamble are multiplexed in the time domain and the data words (**C**$^{(i)}$) of data blocks (**C**) relating to the user data sequence are multiplexed in the frequency domain.

19. The method according to one of claims 1 to 18,
wherein the data blocks (**C**) are obtained from the data signal by means of block coding or by means of permutation.

20. The method according to one of claims 1 to 19,
wherein the data symbols ($c_j^{(i)}$) are modulated onto subcarriers which are orthogonal to each other.

21. A multiplexer (26) adapted to multiplex data words in accordance with the method according to one of claims 1 to 20.

22. A demultiplexer adapted to demultiplex data words multiplexed by the multiplexer of claim 21.

23. A transceiver for wireless communication, comprising at least one of a multiplexer according to claim 21 and a demultiplexer according to claim 22.

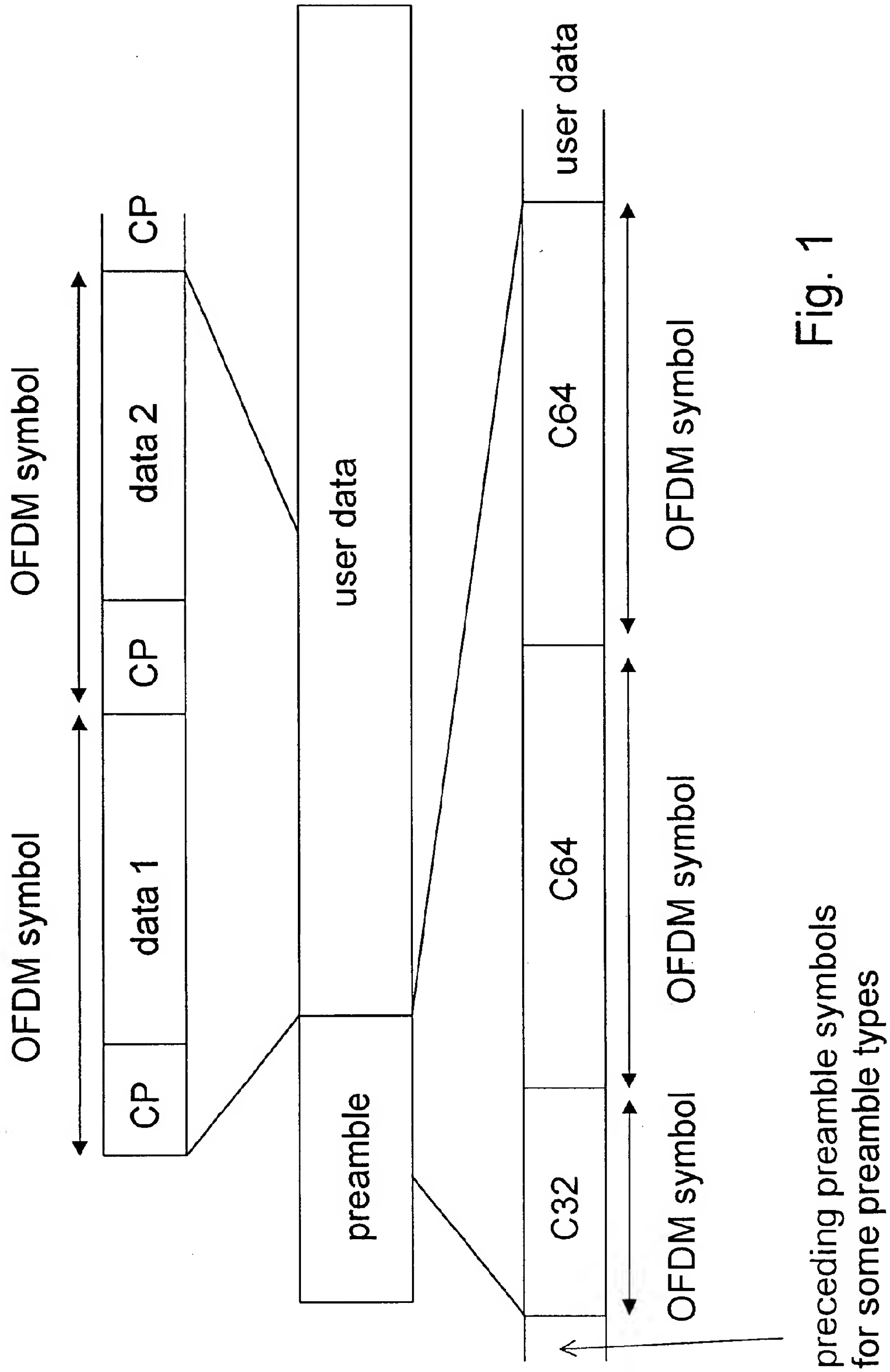
OFDM symbol
OFDM symbol
CP
data 1
CP
data 2
CP
preamble
user data
C32
C64
C64
user data
OFDM symbol
OFDM symbol
OFDM symbol
preceding preamble symbols
for some preamble types

Fig. 1

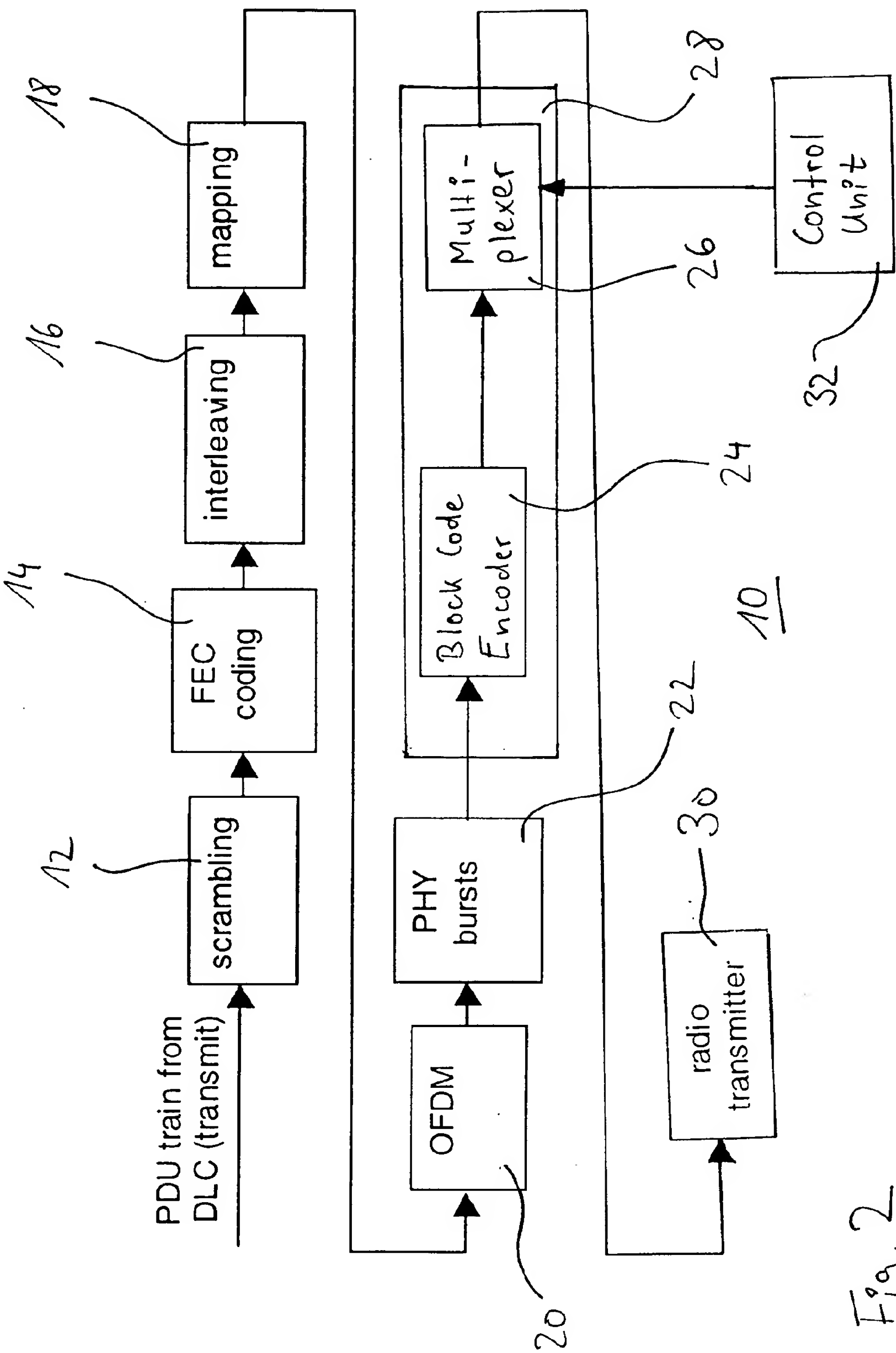
PDU train from DLC (transmit)
scrambling
FEC coding
interleaving
mapping
OFDM
PHY bursts
Block Code Encoder
Multi-plexer
Control Unit
radio transmitter
12
14
16
18
20
22
24
26
28
30
32
10

Fig. 2

modulation scheme	code rate	bit rate
BPSK	1/2	6 Mbps
BPSK	3/4	9 Mbps
QPSK	1/2	12 Mbps
QPSK	3/4	18 Mbps
16-QAM	9/16	27 Mbps
16-QAM	3/4	36 Mbps
64-QAM	3/4	54 Mbps

Fig. 3

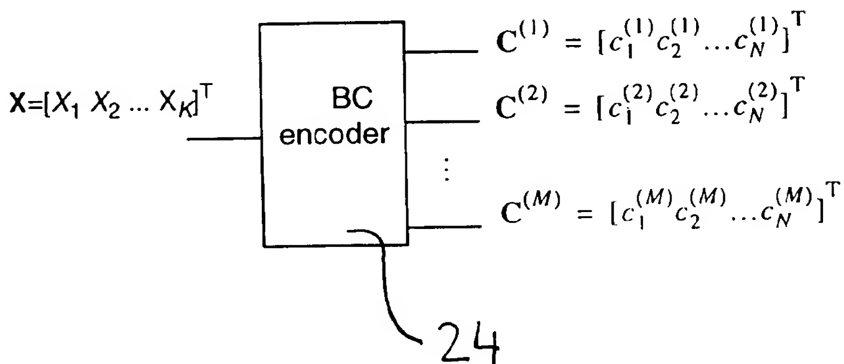

Fig. 4

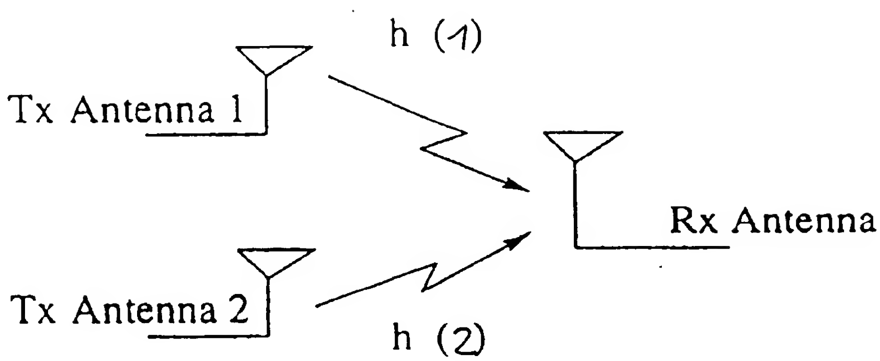

Fig. 5

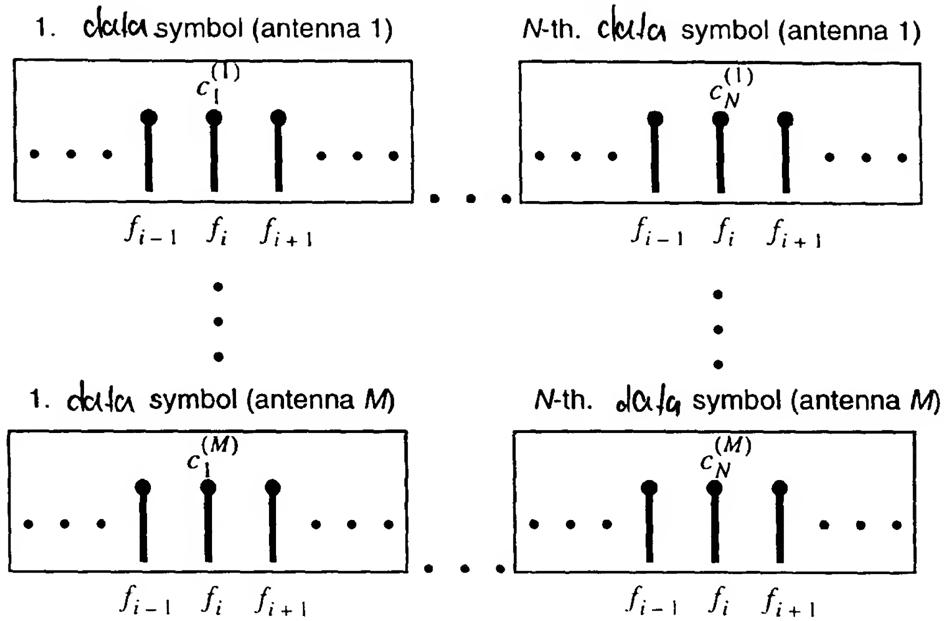
1. data symbol (antenna 1)
N-th. data symbol (antenna 1)
$c_1^{(1)}$
$c_N^{(1)}$
f_{i-1} f_i f_{i+1}
1. data symbol (antenna M)
N-th. data symbol (antenna M)
$c_1^{(M)}$
$c_N^{(M)}$

Fig. 6

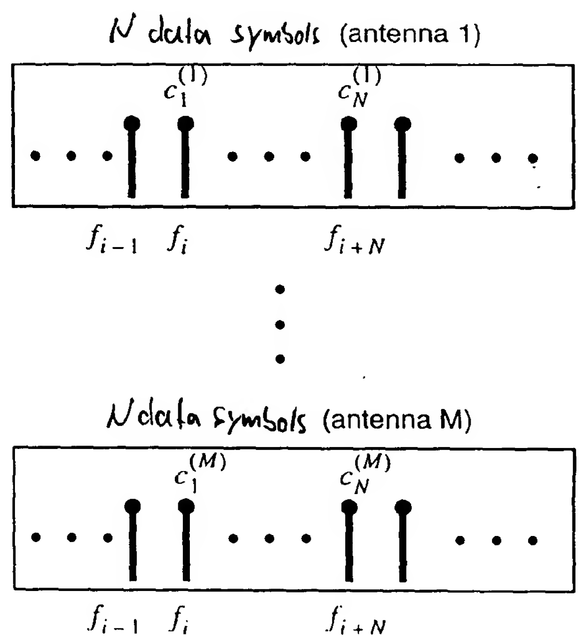
N data symbols (antenna 1)
$c_1^{(1)}$ $c_N^{(1)}$
f_{i-1} f_i f_{i+N}
N data symbols (antenna M)
$c_1^{(M)}$ $c_N^{(M)}$

Fig. 7

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 10 5482

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
D,X	MUDULODU S ET AL: "A transmit diversity scheme for frequency selective fading channels" GLOBECOM '00 - IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE. CONFERENCE RECORD (CAT. NO.00CH37137), PROCEEDINGS OF GLOBAL TELECOMMUNICATIONS CONFERENCE, SAN FRANCISCO, CA, USA, 27 NOV.-1 DEC. 2000, pages 1089-1093 vol.2, XP002172508 2000, IEEE, Piscataway, NJ, USA ISBN: 0-7803-6451-1 * the whole document *	1-23	H04L5/02 H04L1/00 H04B7/06 H04B7/12 H04L1/06 H04L1/04
X	LEE K F ET AL: "A space-frequency transmitter diversity technique for OFDM systems" GLOBECOM '00 - IEEE. GLOBAL TELECOMMUNICATIONS CONFERENCE. CONFERENCE RECORD (CAT. NO.00CH37137), PROCEEDINGS OF GLOBAL TELECOMMUNICATIONS CONFERENCE, SAN FRANCISCO, CA, USA, 27 NOV.-1 DEC. 2000, pages 1473-1477 vol.3, XP002172509 2000, Piscataway, NJ, USA, IEEE, USA ISBN: 0-7803-6451-1 * the whole document *	1-23	
A	ALAMOUTI S M: "A simple transmit diversity technique for wireless communications" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS,IEEE INC. NEW YORK,US, vol. 16, no. 8, October 1998 (1998-10), pages 1451-1458, XP002100058 ISSN: 0733-8716 * the whole document *	1-23	

TECHNICAL FIELDS SEARCHED (Int.Cl.7)
H04L
H04B

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	19 July 2001	Toumpoulidis, T

CATEGORY OF CITED DOCUMENTS

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(54) Multicarrier system with adaptive bit-wise interleaving

(57) The present invention relates to a method for transmitting data streams of users via a transmission path in an OFDM system, whereby the data streams of respective users are transmitted in blocks, frequency hopping according to a predefined frequency hopping pattern for the respective transmission is performed, the block size for each user and within a hopping pattern can vary, and consecutive bits of the data stream to be transmitted are bit-wise interleaved, so that consecutive bits are transmitted in non-adjacent time slots and/or subcarriers within a block according to a predefined interleaving pattern. Thereby, the respective interleaving pattern is made adaptive to the respective frequency hopping pattern and/or the respective block size.

The present invention relates further to a transmitting station (1) for transmitting data streams of users and a receiving station (2) for receiving data streams of users for carrying out this method.

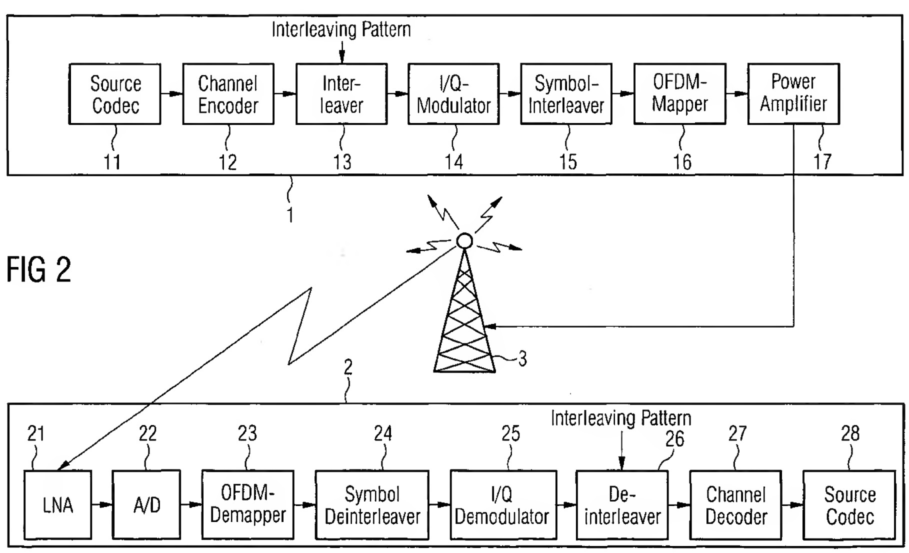

FIG 2

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Description

[0001] The present invention relates to a method for transmitting data in an orthogonal frequency division multiplex system (OFDM System). The present invention relates further to a transmitting station and a receiving station for carrying out this method.

[0002] The so-called orthogonal frequency division multiplex system (OFDM system) is widely used in broadcast systems like Digital Audio Broadcasting (DAB) and Digital Video Broadcasting (DVB) as e.g. described in EN 300 744 V1.2.1 of ETSI (European Telecommunication Standards Institute). This transmission system is also recommended for future wireless communication systems like BRAN (Broadband Radio Access Networks) and HIPERLAN (High Performance Radio Local Area Networks) as described in ETSI TS 101 475 V1.1.1 in order to provide high data rate services. Within these systems, the introduction of a bit-wise interleaver has increased the performance for high-level modulation schemes (e.g. 16QAM, 64QAM, Quadrature Amplitude Modulation), which are required for data transmission at a high-transfer rate.

[0003] A bit-wise interleaver interleaves consecutive and adjacent, respectively, bits of a data stream in a way, that adjacent bits of the data stream are transmitted in non-adjacent time slots and subcarriers, respectively. This has the advantage that associated transfer functions of respective consecutive bits (i.e. subcarrier and/ or time slot number) are uncorrelated. In other words, particularly negative properties like e.g. deep fading of a specific transfer channel, do not take effect on consecutive bits, but rather only on single bits, e.g. of one transmitted symbol.

[0004] In wireless multi-user OFDM systems like e.g. BDMA systems (Band Division Multiple Access), for each user a predefined or fixed number of subcarriers and time slots can be assigned based on the required data rate. This assignment can change pseudo-randomly across the time/frequency-grid of an OFDM transmission path. This assignment is defined as one hopping pattern for one user, where at any one time slot some predefined or fixed subcarriers are allocated.

[0005] Figure 1 illustrates the transmission of data in an OFDM system with frequency hopping, whereby figure 1 shows a hopping pattern (assignment) for one user in a wireless multi-user OFDM system. Thereby, the time axis is divided into time slots of a predefined length and the frequency axis is divided into subcarriers of a predefined bandwidth.

[0006] As seen in figure 1, the data transmission for one user takes place in blocks, whereby each block has a length of a predefined number of time slots and a width of a predefined number of subcarriers. According to a hopping pattern, the (frequency) location, i.e. the frequencies occupied by a respective block within the transmitting path, of each block changes pseudo- randomly.

[0007] As shown in figure 1, the transmission of one user can also take place in more than one block within one period of time slots. This is illustrated in figure 1 as a hatched and a massive block, whereby both blocks belong to the data transmission of one user.

[0008] For each user, the respective corresponding hopping pattern can be repeated within any period of time, e.g. within one frame, one superframe or any other predefined fixed time period. In order to reduce the control burden, the hopping pattern for the respective user is assigned during a link initialisation and establishment phase and it does usually not change before the respective link (i.e. data transmission of one user) is released.

[0009] The block size, i.e. number of time slots and number of subcarriers, can be different for different users, it can also change from block to block (and between the different hopping steps, respectively) within a hopping pattern. These parameters are dependent of the required data rate and the resource management of the transmitting station (base station).

[0010] However, introducing a bit-wise interleaver in such an OFDM system, which performs frequency hopping as described in relation to figure 1, has the disadvantage, that, particularly at a relative small block size, the associated channel transfer functions can not be kept uncorrelated, since the time slot and/or subcarrier distance between the consecutive bits can not been made large enough.

[0011] Thus, due to the limited number of subcarriers at one specific time slot and the limited block size, respectively, it is difficult to implement the bit-wise interleaver according to the prior art in an OFDM frequency hopping system.

[0012] From WO 00/35102 an interleaving/de-interleaving device and a method for a communication system are known. A device for sequentially storing input bit symbols of a given interleaver size in a memory at an address and reading the stored bit symbols from the memory is provided. This known implementation method for an interleaver can be used for example based on CDMA-2000 specification or for other IMT-2000 communication systems. However, it cannot find application for the design of interleaver patterns for multi-user OFDM hopping systems.

[0013] From US 6,125,150 a transmission system using code design for transmission with periodic interleaving is known. Thereby an OFDM transmission system provides a high level of performance on a variety of frequency selective channels by using a code having the characteristics of maximum PPD and maximum PECL. The codes are designed to allow high SNR sub-channels to carry their full potential of information which is then used to compensate for information lost on low SNR sub-channels. According to this known technology error control coding, modulation and interleaver are combined together to obtain better distance characteristics, where some subcarriers may carry more information and another sub-carriers may carry less information

depending on the channel transfer functions.

[0014] It is therefore the object of the present invention to provide a technique for transmitting data streams in an orthogonal frequency division multiplex system (OFDM system), whereby the performance of interleaving and therefore the performance of the transmission is improved.

[0015] The above object is achieved by a method of transmitting data streams of users via a transmission path in a OFDM system according to claim 1.

[0016] This object is further achieved by a transmitting station and a receiving station for carrying out this method according to claims 7 and 8.

[0017] The method for transmitting data streams of users via a transmission path in a OFDM system according to the present invention performs a data transmission. The time axis of the transmission path is divided into time slots The frequency axis of the transmission path is divided into subcarriers. The resource of the transmission path is used by a plurality of users. The data streams of the respective users are transmitted in blocks with a block size of a predefined length of time slots and a predefined number of subcarriers. Frequency hopping according to a predefined frequency hopping pattern for the respective transmission is performed. The frequency hopping pattern for a respective transmission can differ between different users and it also can differ between different times for the same user.

[0018] Further, the frequencies occupied by a respective block within the transmission path vary according to the frequency hopping pattern. The block size for each user within a hopping pattern can also vary. Consecutive bits of the data stream to be transmitted can be bit-wise interleaved such that consecutive bits are transmitted in non-adjacent time slots and/or subcarriers according to a predefined interleaving pattern.

[0019] According to the present invention, the respective interleaving pattern is thereby made adaptive (and can be a function of) to the respective frequency hopping pattern and/or the respective block size.

[0020] The transmitting station according to the present invention for transmitting data streams of users comprises an interleaving means for bit-wise interleaving consecutive bits of data streams according to a predefined interleaving pattern, whereby the interleaving means uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

[0021] The receiving station for receiving data streams of users, which are transmitted according to the above-mentioned method according to the present invention comprises an deinterleaving means for deinterleaving the received data streams into the original bit sequence according to a predefined interleaving pattern, whereby the deinterleaving means uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

[0022] The present invention has the advantage that the performance of high level modulation schemes, like 16QAM, 64QAM or higher, can be improved, since the bit-wise interleaving is made adaptive to the respective data transmission, i.e. the respective hopping pattern and/or block size of the respective transmission. Parasitic characteristics of respective channels are minimised, since e.g. deep fading ideally take only effect on single bits e.g. of one transmitted symbol.

[0023] Thereby, consecutive bits are transmitted within the same block, when the transmitted blocks are large enough, that means, when a block size allows to transmit consecutive bits in the same block so that the associated channel transfer functions keep uncorrelated.

[0024] Consecutive bits can also be transmitted in different blocks according to the interleaving pattern. Advantageously, consecutive bits are transmitted in different blocks, when the block size is very small, so that the associated channel transfer functions for consecutive bits can be kept uncorrelated.

[0025] Further advantageously, the interleaving pattern is made adaptive to the number of time slots of the respective block and/or the interleaving pattern is made adaptive to the number of subcarriers of the respective block.

[0026] In the following description a preferred embodiment of the present invention is explained in more detail in relation to the enclosed drawings, in which

Figure 1 shows an example of data transmission in an OFDM system with frequency hopping,

Figure 2 shows a block diagram of a wireless OFDM system according to the present invention,

Figure 3 shows a diagram of 16QAM and 64QAM mappings and the corresponding bit pattern,

Figure 4 shows one example of the mapping of one symbol into the time-frequency grid of OFDM, and

Figure 5 shows another example of the mapping of one symbol into the time-frequency grid of OFDM.

[0027] Figure 2 shows a schematic diagram of a wireless OFDM system according to the present invention, whereby a block diagram of a transmitting station 1 and a block diagram of a receiving station 2 are depicted.

[0028] The transmitting station 1 according to the present invention comprises a source codec 11 for coding the signals which have to be transmitted (e.g. audio or video signal) into a data stream of a digital signal, and a channel encoder 12 for encoding a data stream e.g. into a frame structure, adding redundancy bits, etc.

[0029] Then, the data stream is adaptively bit-wise interleaved by the interleaver 13 according to the present invention. The pattern for bit-wise interleaving the data stream is thereby made adaptive to predefined param-

eters of the respective transmission like frequency hopping pattern and block size. In other words, the pattern is a function of said parameters. The adaptive interleaving according to the present invention is described later in more detail with reference to Figs. 4 and 5.

[0030] After interleaving the data stream is modulated into symbols, e.g. according to the known I/Q modulation (In-phase/Quadrature modulation), by an I/Q modulator 14 and map into a time/frequency-grid by an OFDM mapper 16. Optionally, the data stream can be symbol-wise interleaved by a symbol interleaver 15 as known from the prior art in order to further improve the transmission performance.

[0031] The OFDM mapper 16 maps the modulated data stream into the time/frequency-grid according to the OFDM transmission system. Further, the OFDM mapper 16 determines the block size and the used frequency hopping pattern, both dependent e.g. upon a given users data rate and resource management in the transmitting station 1.

[0032] Then, the mapped data stream is than amplified by a power amplifier 17 and transmitted via a radio tower 3 over an air-interface to one or plurality of receiving stations.

[0033] The receiving and demodulating of data by the receiving station 2 is carried out in the inverse sequence.

[0034] Thereby, the signal transmitted by the radio tower 3 is received by an antenna comprising a low noise amplifier 21 (LNA). The received signal (comprising the data stream) is analogue/digital converted by an A/D converter 22.

[0035] Complementary to the OFDM mapper 16 of the transmitting station 1, the received signal is demapped by an OFDM demapper 23. The signal is thereby demapped according to the same pattern for mapping the data stream by the OFDM mapper 16 in order to reconstruct the original data stream.

[0036] If the signal is symbol-wise interleaved by the transmitting station 1, the signal has to be symbol-wise deinterleaved by a symbol deinterleaver 24.

[0037] After I/Q demodulation of the demapped data stream by the I/Q demodulator 25, the data stream is bit-wise deinterleaved by the deinterleaver 26. Thereby, the pattern for deinterleaving the data stream is made adaptive to the hopping pattern for mapping/demapping the signal; the pattern for bit-wise deinterleaving is similar to the interleaving pattern used by the transmitter 1 in order to get the original data stream.

[0038] Channel decoding and source decoding is performed by a channel decoder 27 and source decoder 28 similar to the source coding and channel coding of the transmitting station 1.

[0039] Figure 3 shows the principles of QAM (Quadrature Amplitude Modulation) based on the example of 16QAM and 64QAM.

[0040] For QAM the information is transmitted with an in-phase and a quadrature component Q. Thus, the carrier comprises respective to the information, which are transmitted, an in-phase (I) and an quadrature (Q) component. Thereby, dependent on the modulation scheme (e.g. 16QAM or 64QAM), one transmitted symbol carries 4 bits at 16QAM (respectively 2 bits for I- and Q-channel) and 6 bit at 64QAM (respectively 3 bits for I- and Q-channel) pursuant to the scheme as shown in the respective coordinate system of 16QAM and 64QAM; the bit order is termed by I1, Q1, I2, Q2 and I1, Q1, I2, Q2, I3, Q3, respectively.

[0041] Thereby, the high priority bits are I1 and Q1. For 16QAM, the low priority bits are I2 and Q2, for 64QAM, the low priority bits are I3 and Q3. At the example of 64QAM (encircled symbol 000011) it is illustrated, that the high priority bits are less susceptible against interferences than the low priority bits. If, e.g., this symbol is interfered, it could be decoded wrongly as an adjacent symbol, e.g. as 000010, 000111, 001011 or 000001. It is seen, that the high priority bits are always the same, namely 00. Thus, the high priority bis are more protected against interferences than the low priority bits.

[0042] Figure 4 shows one example of a pattern for mapping a data stream into a time/frequency-grid by the OFDM mapper 16 shown in figure 2.

[0043] In this example, the hopping pattern for one user is shown. Thereby, each use is assigned two blocks (shown as hatched and solid blocks). Since the respective blocks are very small, i.e. low number of subcarriers in this example, consecutive bits of one symbol, e.g. I1, Q1, I2, Q2 are transmitted in different blocks, interleaved according to a bit-wise interleaving pattern A.

[0044] Figure 5 shows a different hopping pattern for mapping a data stream into the time/frequency-grid.

[0045] Thereby, the block size differs between the single hopping steps. E.g. in the first block and the third block are transmitted two consecutive bits (respectively I1, Q1 and I2, Q2), since the block size is large enough. In this case, bit-wise the interleaving happens according to an interleaving pattern B.

[0046] Thus, the design rule of a bit-wise interleaver 13 as shown in figure 1 is as follows:

- Adjacent coded bits from channel encoder are mapped onto non-adjacent subcarriers or non-adjacent time slots. The frequency separation (distance) of the chosen subcarriers or the time separation of the chosen time slot has to be far enough in order to keep the associated channel transfer functions uncorrelated.

- Adjacent coded bits from channel encoder are mapped alternatively on high or low priority bits. By this way, long runs of low reliability bits are avoided.

- The coded bits are placed at all available subcarriers and time slots on OFDM time/frequency-grid within the depth, i.e. the time/frequency/block dis-

tance between consecutive bits, of the interleaver are used.

- The bit-wise interleaver pattern is made adaptive to the hopping pattern in order to achieve better system performance.

[0047] The present invention has the advantage, that the performance of high level modulation schemes, like 16QAM, 64QAM or higher, can be improved, since the bit-wise interleaving is made adaptive to the respective data transmission, i.e. the respective hopping pattern and/or block size of the respective transmission. Parasitic characteristics of respective channels are minimised, since e.g. deep fading ideally take only effect on single bits e.g. of one transmitted symbol.

[0048] According to invention therefore a new design rule is proposed for a bit-wise interleaver for multi-user OFDM hopping systems. Instead of placing data bits belonging to one I/Q symbol or adjacent symbols on different sub-carriers at the same timeslot, data bits belonging to one I/Q symbol or adjacent symbols can be placed at different timeslots or at different blocks within the bit-wise interleaver.

[0049] Furthermore, in multi-user OFDM hopping systems, where each user can be assigned different hopping patterns (depending upon a given user's data rate and resource management in the base station) at different times, it is proposed that the practical bit-wise interleaver pattern to be used for each user is variable and depends upon its assigned hopping pattern. In this way the optimal performance can be obtained.

[0050] Each sub-carrier thereby carries the same size of information. By using a bit-wise interleaver, the bits belonging to one symbol are interleaved. Therefore, the good performance can be achieved for error control coding. Therefore, the use of a bit-wise interleaver can be enabled for multi-user OFDM hopping systems.

Claims

1. Method for transmitting data streams of users via a transmission path in an OFDM system, whereby
the time axis of the transmission path is divided into time slots,
the frequency axis of the transmission path is divided into subcarriers,
the transmission path is used by a plurality of users,
the data streams of the respective users are transmitted in blocks with a block size of a predefined length of time slots and a predefined number of subcarriers,
frequency hopping according to a predefined frequency hopping pattern for the respective transmission is performed, whereby the frequency hopping pattern for respective transmission can differ between different users and differ between different times for the same user,
the frequencies occupied by a respective block within the transmission path vary according to the frequency hopping pattern,
the block size for each user and within a hopping pattern can vary, and
consecutive bits of the data stream to be transmitted are bit-wise interleaved, so that consecutive bits are transmitted in non-adjacent time slots and/or subcarriers according to a predefined interleaving pattern,
characterised in
that the respective interleaving pattern is made adaptive to the respective frequency hopping pattern and/or the respective block size.

2. Method according to claim 1,
characterised in
that consecutive bits are transmitted within the same block, in case the block size is large.

3. Method according to claim 1,
characterised in
that consecutive bits are transmitted in different blocks according to the interleaving pattern.

4. Method according to claim 3,
characterised in
that consecutive bits are transmitted in different blocks, in case the block size is small.

5. Method according to one of the claims 1 to 4,
characterised in
that the interleaving pattern is made adaptive to the number of time slots of the respective block.

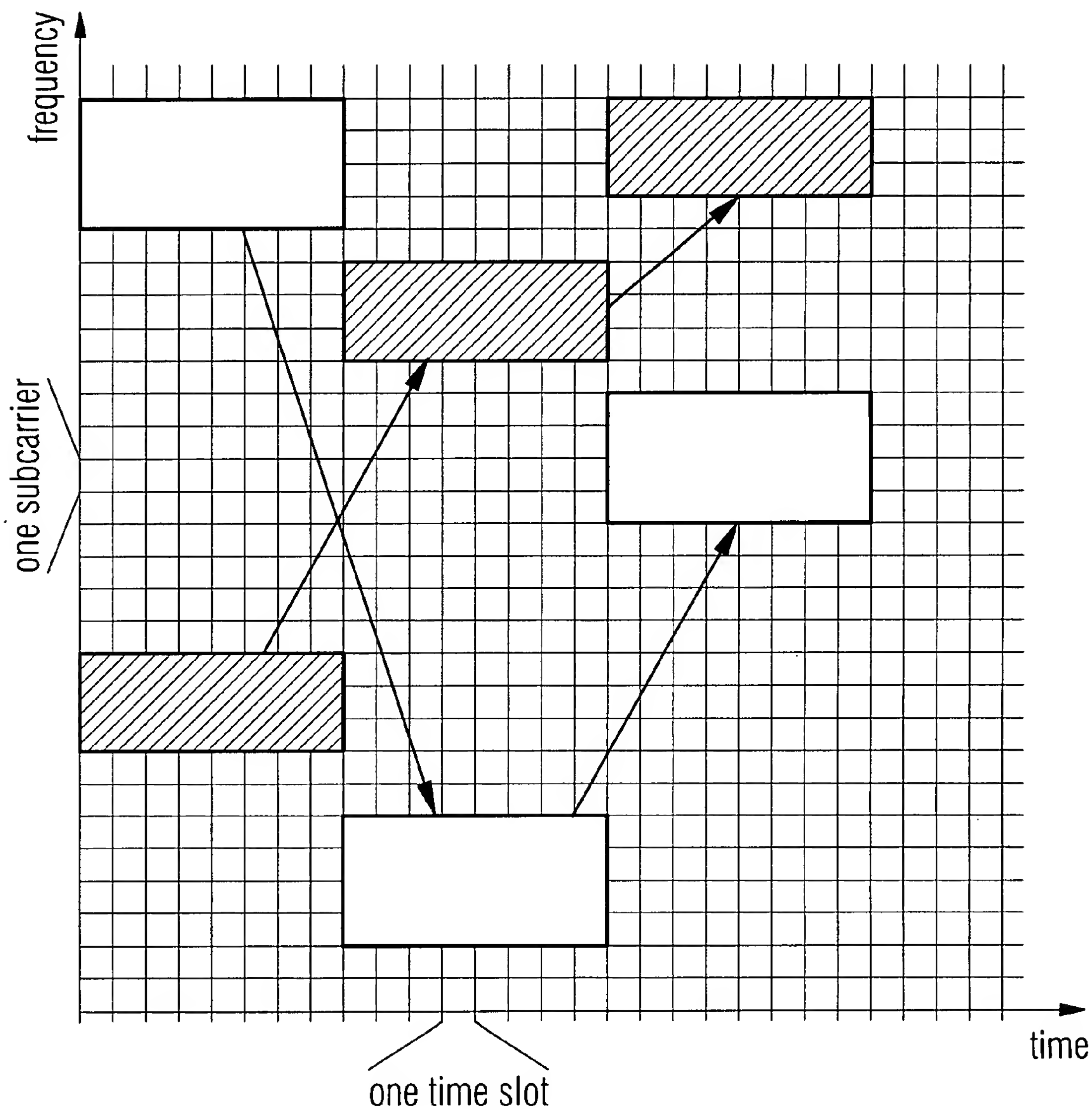
6. Method according to one of the claims 1 to 5,
characterised in
that interleaving pattern is made adaptive to the number of subcarriers of the respective block.

7. Transmitting station (1) for transmitting data streams of users by using a method according to anyone of the claims 1 to 6,
characterised by
an interleaving means (13) for bit-wise interleaving consecutive bits of data streams according to a predefined interleaving pattern, whereby the interleaving means (13) uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

8. Receiving station (2) for receiving data streams of users, which are transmitted using a method for transmitting data streams according to anyone of the claims 1 to 6,
characterised by
an deinterleaving means for deinterleaving the re-

ceived data streams into the original bit sequence according to a predefined interleaving pattern, whereby the deinterleaving means uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

FIG 1



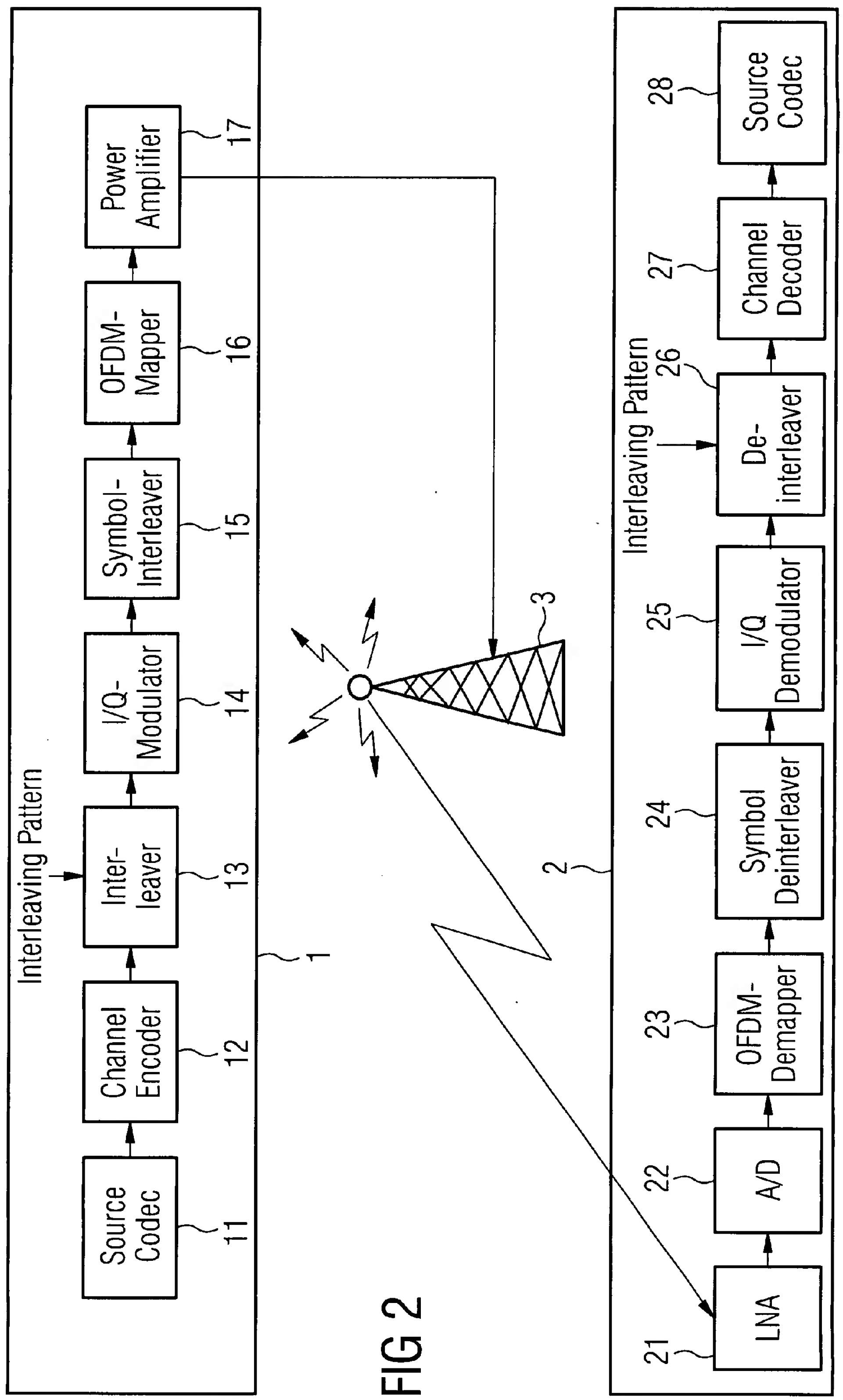

Interleaving Pattern
Source Codec
11
Channel Encoder
12
Inter-leaver
13
I/Q-Modulator
14
Symbol-Interleaver
15
OFDM-Mapper
16
Power Amplifier
17
1
3
2
Interleaving Pattern
LNA
21
A/D
22
OFDM-Demapper
23
Symbol Deinterleaver
24
I/Q Demodulator
25
De-interleaver
26
Channel Decoder
27
Source Codec
28

FIG 2

FIG 3

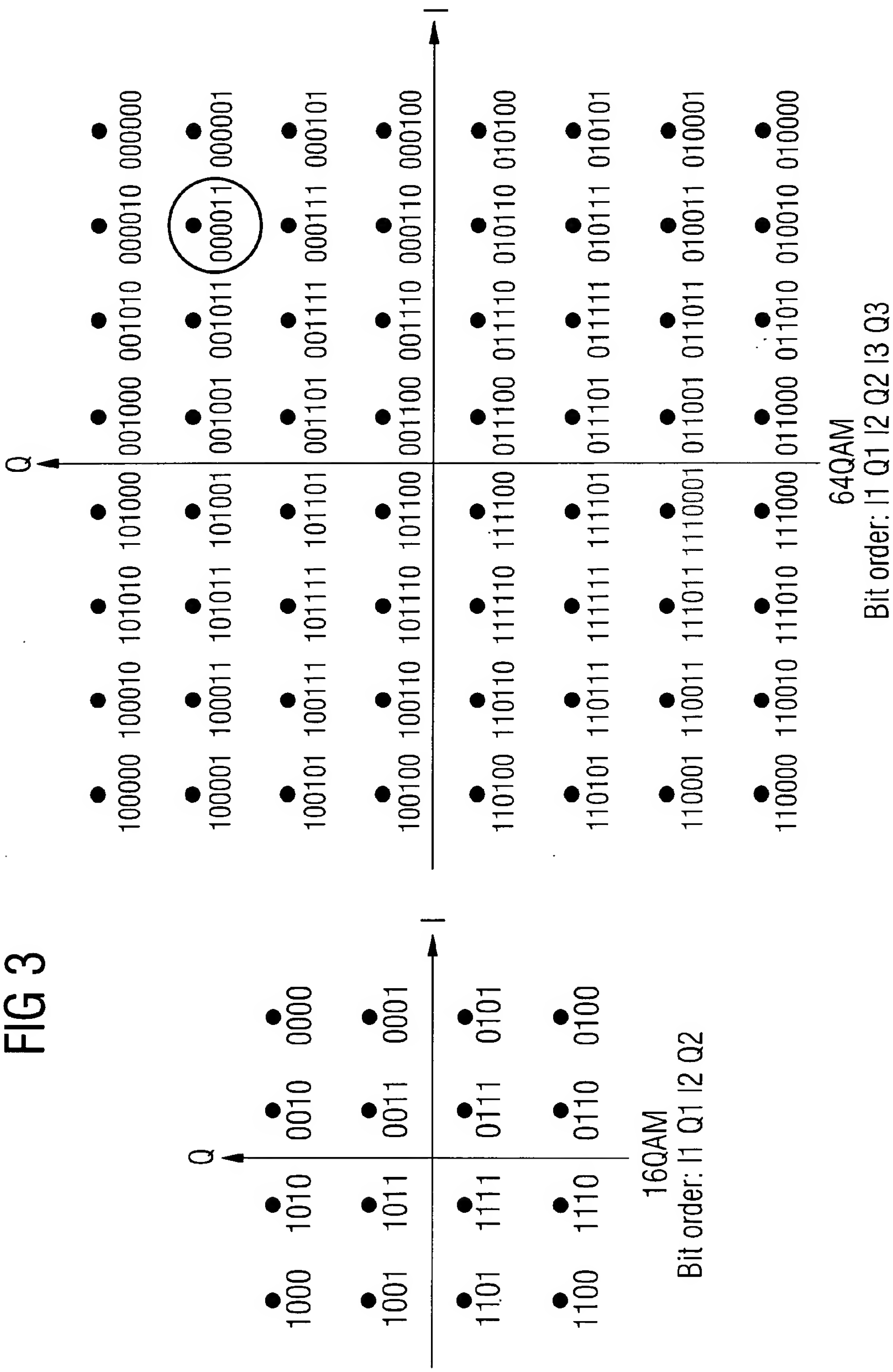

Q
I
1000 1010 0010 0000
1001 1011 0011 0001
1101 1111 0111 0101
1100 1110 0110 0100
16QAM
Bit order: I1 Q1 I2 Q2
100000 100010 101010 101000 001000 001010 000010 000000
100001 100011 101011 101001 001001 001011 000011 000001
100101 100111 101111 101101 001101 001111 000111 000101
100100 100110 101110 101100 001100 001110 000110 000100
110100 110110 111110 111100 011100 011110 010110 010100
110101 110111 111111 111101 011101 011111 010111 010101
110001 110011 111011 1110001 011001 011011 010011 010001
110000 110010 111010 111000 011000 011010 010010 010000
64QAM
Bit order: I1 Q1 I2 Q2 I3 Q3

FIG 4
Interleaving Pattern A

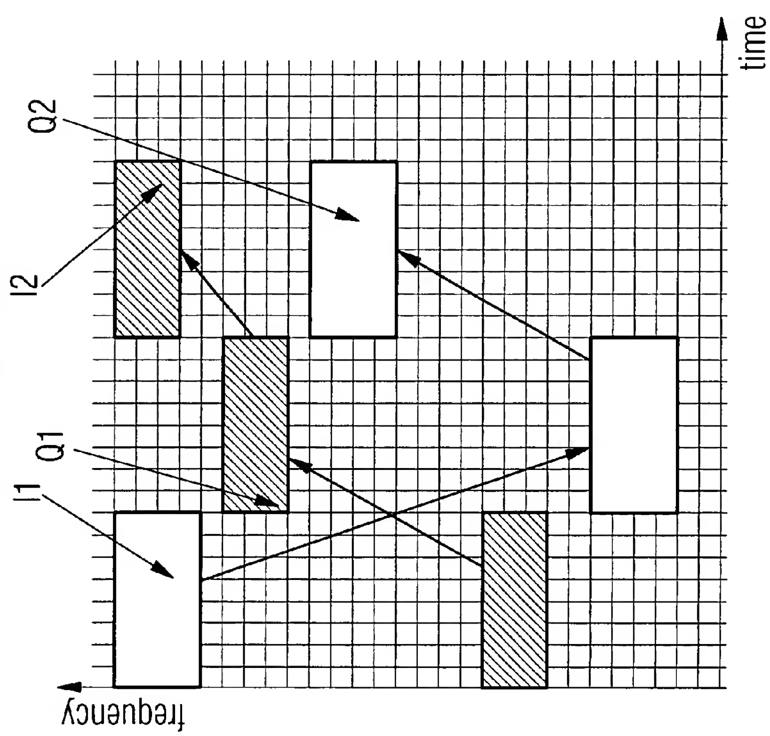
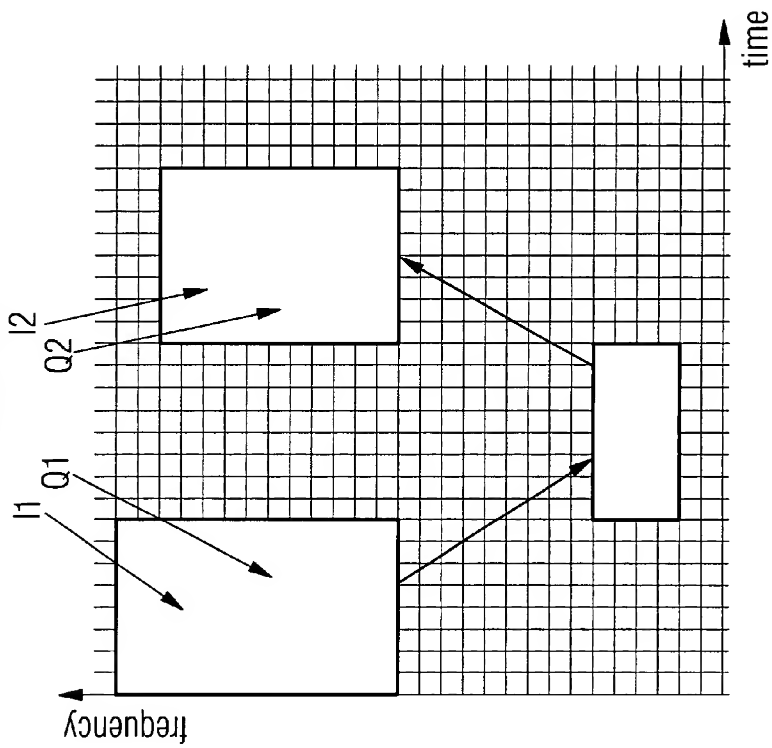


FIG 5
Interleaving Pattern B



European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 11 4041

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Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X Y	US 5 425 051 A (MAHANY RONALD L) 13 June 1995 (1995-06-13) * abstract * * column 2, line 27 - line 36 * * column 2, line 50 - line 64 * * column 14, line 56 - line 60 * * column 16, line 55 - line 60 * * column 17, line 38 - line 45 * * column 18, line 38 - line 56; claim 20 *	7,8 1	H04L27/26
D,Y A	US 6 125 150 A (CIOFFI JOHN M ET AL) 26 September 2000 (2000-09-26) * abstract * * column 4, line 61 - column 5, line 4 * * column 6, line 22 - line 28 * * claims 8,12,14 *	1 7,8	
A	US RE36430 E (HALBERT-LASSALLE ROSELYN ET AL) 7 December 1999 (1999-12-07) * abstract; figure 5 * * column 3, line 17 - line 37 * * column 8, line 30 - line 36 * * column 8, line 50 - line 56 *	1,7,8	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04L H03M
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The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	17 October 2001	Papantoniou, A

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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17-10-2001

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For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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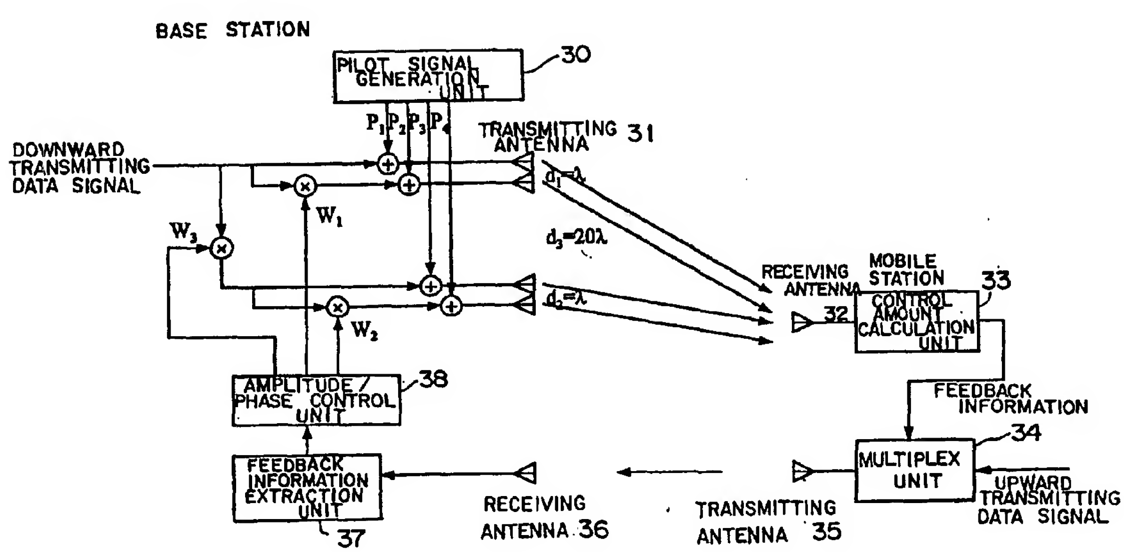
(54) TRANSMISSION DIVERSITY COMMUNICATION DEVICE

(57) The plurality of antennas of a base station used for transmitting diversity are divided into groups. Each antenna is located so that signals transmitted from antennas in the same group have a high fading correlation. Each antenna group is spaced so that a fading correlation between the groups may become low. Since signals transmitted from an antennas in the same group have high fading correlation, such signals suffer little from fading fluctuations and a low control speed is acceptable. However, control between the groups must be exercised at a high speed. Therefore, a mobile station that receives the signals of the base station feeds back feedback information for controlling fading fluctuations between the groups and information within a group to the base station at a high transfer rate and at a low transfer rate, respectively.



F I G. 4

EP 1 315 311 A1

Description

Technical field

[0001] The present invention relates to a transmitting diversity communications apparatus.

Background Art

[0002] Transmitting diversity in W-CDMA, which is the third-generation mobile communications system, adopts a method using two transmitting antennas.

[0003] Fig. 1 shows an example configuration of a transmitting diversity system using two transmitting antennas.

[0004] Mutually orthogonal pilot patterns P_1 and P_2 are transmitted from two transmitting antennas 1 and 2, respectively, as pilot signals, and channel impulse response vectors $\underline{h}_1$ and $\underline{h}_2$ from each antenna of a base station up to the receiving antenna of a mobile station are estimated by correlating each known pilot pattern to an incoming pilot on the receiving side of the mobile station.

[0005] A control amount calculation unit 10 calculates and quantizes the amplitude/phase control vector (weight vector) $\underline{w}=[w_1, w_2]$ of each transmitting antenna of the base station that maximizes power P expressed by the following equation (1) using these channel estimation values. Then, a multiplex unit 11 multiplexes the quantized weight vectors with an uplink channel signal as feedback information and transmits the signal to the base station. However, since there is no need to transmit both values w_1 and w_2, it is acceptable to transmit only value w_2 obtained by assigning $w_1=1$.

$$P = \underline{w}^H H^H H \underline{w} \tag{1}$$

$$H = [\underline{h}_1, \underline{h}_2] \tag{2}$$

[0006] In equation (2), $\underline{h}_1$ and $\underline{h}_2$ are the channel impulse response vectors from the transmitting antennas 1 and 2, respectively, and the superscript H on H^H and w^H indicates the Hermitian conjugation of H and w, respectively. If an impulse response length is assumed to be L, the channel impulse response vector is expressed as follows.

$$\underline{h}_i = [h_{i1}, h_{i2}, \ldots, h_{iL}] \tag{3}$$

[0007] Therefore, in the case of two transmitting antennas, equation (1) is calculated based on the following algebraic calculation.

$$H = \begin{bmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \\ \vdots & \vdots \end{bmatrix} \quad , \quad \underline{w} = [w_1, w_2]^T \quad , therefore \quad Hw = \begin{bmatrix} h_{11}w_1 + h_{21}w_2 \\ h_{12}w_1 + h_{21}w_2 \\ \vdots \end{bmatrix}$$

[0008] At the time of handover, weight vector $\underline{w}$ that maximizes the following equation is calculated instead of equation (1).

$$P = \underline{w}^H (H^H{}_1 H_1 + H^H{}_2 H_2 + \ldots) w \tag{4}$$

[0009] In equation (4), H_K is a channel impulse response signal from the k-th base station.

[0010] Then, the feedback information extraction unit 12 on the transmitting side extracts w_2 (in this case, $w_1=1$ is assumed) transmitted from a mobile station, from an incoming signal and an amplitude/phase control unit 13 multiplies a data signal to be transmitted from the transmitting antenna 2 by W_2. In this way, the degradation of both the amplitude and phase of signals received from the transmitting antennas 1 and 2 that are received on the receiving side are corrected in advance and are transmitted from the transmitting side.

[0011] In W-CDMA, two methods are stipulated: mode 1 for quantizing weight coefficient w_2 into one bit and mode 2 for quantizing w_2 into four bits. In mode 1, since control is exercised by transmitting one bit of feedback information for each slot, control speed is high. However, since quantization is rough, accurate control is impossible. In mode 2, since control is exercised by transmitting four bits of information, more accurate control is possible. However, in mode 2, since only one bit can be transmitted for each slot and feedback information of one word is transmitted for every four slots, control cannot track fading in the case of a high fading frequency, and amplitude/phase characteristics degrade. As described above, if the signal transfer rate of an uplink channel from a mobile station to a transmitting station, for transmitting feedback information is restricted, control accuracy and fading track speed have an inverse relationship.

[0012] The Release-99 specification of W-CDMA standard does not take into consideration a case where more than two transmitting antennas are used so as to avoid the degradation of uplink channel transmission efficiency due to feedback information transmission. However, if the increase of feedback information or the degradation of update speed is allowed, the number of antennas can also be increased to three or more. In particular, currently a case where four transmitting antennas are used is being extensively researched and developed.

[0013] If a closed-loop transmitting diversity system is applied to the radio base station of a cellular mobile communications system, a signal from each transmitting antenna independently suffers from fading, and ideally the same phase combination is performed at the antenna position of the mobile station. Therefore, a diversity gain corresponding to the number of transmitting antennas can be obtained and the gain can also be improved by the combination. Accordingly, the receiving characteristic is improved and the number of users accommodated in one cell can also be increased. "Ideally" means a case where there is neither transmission error of feedback information, control delay, channel response estimation error nor quantization error of a control amount. In reality, the characteristic degrades due to these factors compared with that of the ideal case.

[0014] In order to obtain a diversity gain corresponding to the number of antennas, antenna spacings (the distances between antennas) must be wide so that fading correlation may become sufficiently low. Generally, in order to suppress fading correlation to a sufficiently low level in the radio base station of a cellular mobile communications system, antenna spacings must be approximately 20 wavelengths. Since one wavelength is approximately 15cm in a 2GHz band, antennas must be installed approximately 3 meters apart. Therefore, if the number of transmitting antennas increases, an area needed to install antennas becomes wide and it becomes difficult to install antennas on the roof of a building and the like, which is a problem. Diversity gain is saturated as the number of transmitting antennas increases. Therefore, when the number of transmitting antennas reaches a specific value, the diversity gain cannot be improved any further even if the number of transmitting antennas is further increased.

[0015] When the number of transmitting antennas is increased, an amount of information to be fed back increases since feedback information must be transmitted to each antenna. Therefore, in that case, the transmission efficiency of an uplink channel degrades due to feedback information transmission or the control of transmitting diversity cannot track high-speed fading. As a result, the characteristic degrades, which is another problem.

Disclosure of Invention

[0016] An object of the present invention is to provide a transmitting diversity communications apparatus for suppressing the increase of uplink feedback information if the number of transmitting antennas is increased, suppressing the degradation of a characteristic in the case of a high fading frequency and requiring a small antenna installation space in the base-station.

[0017] The transmitting diversity communications apparatus of the present invention includes a transmitting diversity base station for controlling transmitting signals, according to information from a mobile station. The transmitting diversity communications apparatus comprises an antenna unit composed of a plurality of antenna groups, each consisting of a plurality of antennas, located close to each other so that the fading correlation between the antennas in the same group is high and groups are located apart from one another so that the fading correlation between the groups is low, and a control unit receiving both the first control information about intra-group antenna control with a low transfer rate that is transmitted from a mobile station and the second control information about inter-group antenna control and controlling the phase of a signal transmitted by the antenna unit.

[0018] According to the present invention, if signal control is applied to a closed-loop transmitting diversity system by the same method as in the conventional case where two transmitting antennas are used, by increasing the number of transmitting antennas, the tracking of fading fluctuations and transmitting-signal control performance can be prevented from degrading due to the increase of an amount of information to be transmitted from a mobile station to a base station.

[0019] In particular, according to the present invention, since the antenna unit of a base station is composed of a plurality of antenna groups each consisting of a plurality of antennas, and each intra-group antenna and each antenna group is set so that fading correlation is high within a group and so that fading correlation is low between groups,

respectively, only transmitting-signal control information between groups must be transmitted at a high speed from a mobile station to a base station and transmitting-signal control information within a group can be relatively slow. Therefore, transmitting diversity performance can be improved by effectively utilizing the limited transfer rate of an upward line from a mobile station to the base station.

Brief Description of Drawings

[0020]

Fig. 1 shows an example configuration of a transmitting diversity system using two transmitting antennas.
Fig. 2 shows the system configuration of the present invention.
Fig. 3 shows an example configuration of transmitting antennas of a base station according to the preferred embodiment of the present invention.
Fig. 4 shows the configuration of one preferred embodiment of the present invention.
Fig. 5 shows an example of a downlink pilot signal pattern in the preferred embodiment.
Fig. 6 shows both an example configuration of a base station transmitting antennas and antenna control information according to the preferred embodiment.
Fig. 7 shows an envelope correlation coefficient obtained when the angle dispersion $\Delta\phi$ of an input signal observed at a base station in a macro-cell environment is approximately 3.
Fig. 8 shows an example of the transmission format of feedback information in the preferred embodiment (No. 1) .
Fig. 9 shows an example of the transmission format of feedback information in the preferred embodiment (No. 2).
Fig. 10 shows an example of the transmission format of feedback information in the preferred embodiment (No. 3).
Fig. 11 shows an example of the transmission format of feedback information in the preferred embodiment (No. 4).
Fig. 12 shows an example configuration of a mobile station for transmitting feedback information to a base station according to the formats shown in Figs. 8 through 11.
Fig. 13 shows an example configuration of a base station in the second preferred embodiment of the present invention.
Fig. 14 shows an antenna phase difference control method within a group in the second preferred embodiment.
Fig. 15 shows the configuration of the third preferred embodiment of the present invention.

Best Mode for Carrying Out the Invention

[0021] The present invention relates to a closed-loop transmitting diversity method according to which the radio base station of a cellular mobile communications system is provided with a plurality of antennaslements, both different amplitude and phase control are exercised over the same transmitting data, according to feedback information from a mobile station and a plurality of pieces of data are transmitted using different antennas. On the mobile station side, the amplitude/phase control amounts are determined using a downward pilot signal; feedback information indicating the amplitude/phase control amounts are multiplexed with an uplink channel signal; and the data is transmitted to the base station.

[0022] Fig. 2 shows the system configuration of the present invention.

[0023] The pilot signal generation unit 20 of a base station generates N mutually orthogonal pilot signals $P_1(t)$, $P_2(t)$, ..., $P_N(t)$ and the pilot signals are transmitted using different antennas. N is the number of transmitting antennas. The following relationship is established between these pilot signals.

$$\int P_i(t)P_j(t)dt = 0 \quad (i \neq j)$$

[0024] Each pilot signal suffers from both amplitude and phase fluctuations due to fading, and a signal obtained by combining these pilot signals is transmitted to the receiving antenna 22 of a mobile station. The receiver of the mobile station estimates the channel impulse response vectors h_1, h_2, ..., h_N of each pilot signal by calculating the correlation between the incoming pilot signal and each of $P_1(t)$, $P_2(t)$, ..., $P_N(t)$.

[0025] A control amount calculation unit 23 calculates and quantizes the amplitude/phase control vector (weight vector) $\underline{w}=[w_1, w_2, ..., w_N]^T$ of each transmitting antenna of the base station that maximizes power P expressed by equation (5)(the same as equation (1)) using these channel impulse response vectors. A multiplex unit 24 multiplexes the quantized vector with an upward channel signal as feedback information and transmits the signal to the base station side. However, in this case it is acceptable to transmit values w_2, w_3, ..., w_N obtained by assigning $w_1=1$.

$$P = \underline{w}^H H^H H \underline{w} \tag{5}$$

$$H = [\underline{h}_1, \underline{h}_2, .., \underline{h}_N] \tag{6}$$

[0026] In equation (6), $\underline{h}_1$ is a channel impulse response vector from transmitting antenna i. If an impulse response length is assumed to be L, $\underline{h}_1$ is expressed as follows.

$$\underline{h}_1 = [h_{i1}, h_{i2}, ..., h_{iL}]^T \tag{7}$$

[0027] At the time of hand-over, weight vector w that maximizes the following equation is calculated instead of equation (5).

$$P = \underline{w}^H (H_1^H H_1 + H_2^H H_2 + ...) \underline{w} \tag{8}$$

[0028] In equation (8), H_k is a channel impulse response signal from the k-th base station, and is the same as H_k in equation (4).

[0029] The multiplex unit 24 of the mobile station multiplexes the weight vector obtained in this way with an upward transmitting data signal and the vector is transmitted to the receiving antenna of the base station. In the base station, a feedback information extraction unit 25 extracts the feedback information received by a receiving antenna, and an amplitude/phase control unit 26 controls both the amplitude and phase transmitted from each transmitting antenna using a weight vector included in the feedback information. When the base station transmits a signal, both the amplitude and phase of which have been controlled from a transmitting antenna 21, the mobile station receives the signal as if the fluctuations due to just fading of both the amplitude and phase were compensated for. Therefore, optimal reception is possible. Since fading changes as time elapses, both the generation and transmission of feedback information must happen in real time. However, since both the transmission format and transfer rate of an uplink data signal from a mobile station to a base station is predetermined, it takes too much time to transmit a lot of information. Therefore, the control cannot track the fading fluctuations. In order to track the fading fluctuations, the transmission rate of feedback information must be high. However, since the transmission rate of an uplink control channel is limited, if a plurality of pieces of new information are sequentially transmitted in a short cycle in order to control transmitting diversity at a high speed, an amount of information included in one time transmission becomes small (quantization becomes rough), and highly accurate control becomes impossible.

[0030] In the preferred embodiment of the present invention, each coefficient value in a weight vector is calculated and fed back in a different cycle instead of calculating and feeding back a signal transmitted from each antenna in the same cycle.

[0031] The details are described below.

[0032] Fig. 3 shows an example configuration of the transmitting antennas of a base station according to the preferred embodiment of the present invention.

[0033] As shown in Fig. 3, in a base station, transmitting antennas compose a plurality of groups, each consisting of a plurality of antennas. Transmitting antennas in the same group are located close to one another so that the fading correlation between the antennas is high and groups are installed apart from one another so that the fading correlation between the groups is low. Fading correlation is a numeric value indicating how similarly two signals transmitted from different antennas fade when the signals are received on a receiving side. Doppler effect and the like cause fading by reflection on buildings and mobile objects. Therefore, if a plurality of antennas transmitting signals are located close to one another, it follows that a mobile station receives the respective signals through similar routes. Accordingly, the signals suffer from similar fading. In such a case, it is said that the fading correlation between the signals is high. If a plurality of antenna transmitting signals are located apart from one another, it follows that the respective signals take different routes to a mobile station receives the signals. Therefore, the signals fade differently and then are received by the mobile station. In such a case, it is said that the fading correlation between the signals is low.

[0034] In a mobile station, an antenna control amount between groups is calculated in a shorter cycle than that of the antenna control amount within a group, and is transmitted to a base station side as feedback information. Signals from base-station transmitting antennas in the same group have a high fading correlation; the signals suffer from almost the same fading, but the signals each have a phase difference depending on the angle at which the signals reach the receiving antenna of the mobile station. Therefore, each channel response estimation value estimated using the signals

from the plurality of base-station transmitting antennas in the same group has a phase difference that depends on the angle of the mobile station against the base station. Although these values change as the mobile station travels, the values change slowly compared with fading fluctuations. One antenna in each group is designated as a reference antenna, and each of the control amounts of antennas other than the reference antenna in the relevant group is normalized by the control amount of this reference antenna (each relative value calculated using the control amount of this reference antenna as a reference is used). This normalized antenna control amount in the group changes slowly as the mobile station travels. Therefore, the control cycle can be made relatively long.

[0035] However, since respective signals from base-station transmitting antennas belonging to different groups have a low fading correlation, the signals fade differently and independently by the time they reach the receiving antenna of the mobile station. Therefore, respective channel response estimation values (channel impulse response vector) estimated using respective signals from respective reference values belonging to different groups change quickly due to respective independent fading fluctuations. An antenna control amount obtained by normalizing the reference antenna control amount of one specific group by the reference antenna control amount of another group is defined as an inter-group antenna control amount. Since each inter-group antenna control amount changes quickly due to each independent fading fluctuation, in order to accurately control antennas, the control must be exercised in a short cycle.

[0036] The mobile station must recognize which signal comes from which group. However, it is sufficient to relate each antenna to each pilot signal transmitted from the antenna in advance. Since pilots are mutually orthogonal to one another, a receiving side can accurately recognize from which antenna the signal is transmitted by checking the pilot signal.

[0037] Both the inter-group antenna control amount $F_{1,m}$ and intra-group antenna control amount $G_{m,k}$ shown in Fig. 3 are calculated as follows. In the description given above, N, M and K=N/M are the total number of antennas, the number of antenna groups and the number of antennas in each group, respectively. * represents complex conjugation.

Overall reference antenna: Antenna #1
Intra-group reference antenna: Antenna # ((m-1) K+1) (m-1, ..., M)

$$F_{1.m} = \frac{w_{(m-1)K+1}}{w_1} \quad (m = 1,...,M) \tag{9}$$

$$G_{m,k} = \frac{w_{(m-1)K+k+1}}{w_{(m-1)K+1}} \quad (m = 1,...,M,k = 1,...K) \tag{10}$$

[0038] Since fading correlation is high within a group, $|G_{m,k}| = 1$ can be assigned. Specifically, it can be considered that the change due to fading is small within a group, and it is sufficient to take into consideration only change in phase. In order to keep total transmission power constant (=1.0), $F_{1,m}$ must be normalized as follows.

$$F'_{1,m} = \frac{F_{1,m}}{\sqrt{\frac{1}{KM}\sum_{j=1}^{M}\left|F_{1,j}\right|^2}}$$

[0039] Next, the fluctuation rate of fading is described.

[0040] Fading fluctuation rate is expressed by Doppler frequency.

$$f_d = \frac{v}{\lambda}$$

[0041] In the equation described above, v is the travel speed of a mobile station and λ is the carrier wavelength. For example, if a carrier frequency is 2GHz and the travel speed of a mobile station is 60km/h, f_d becomes approximately 111Hz. However, the angle of arrival of an incoming wave changes as the mobile station travels. For example, if the mobile station travels at a speed of 200km/h at a place 200 meters ahead, the input angle changes by approximately 15 degrees per second. In this way, the fading fluctuation rate is higher by several tens of times to several hundreds of times than the fluctuation rate of an input angle. According to W-CDMA standards, a slot length is 666.7μs and the

update speed of feedback information is 1500Hz. Therefore, if information about fading is not updated for each slot, a track characteristic degrades. However, there is no need to feedback information about input angle for each slot. For example, there will be no problem if information is updated for every 15 slots (=one frame) .

[0042] By utilizing the difference in the fluctuation rates of the control information described above, a feedback amount of information can be reduced without performance degradation. Specifically, an inter-group antenna control amount changing at a high speed is updated and fed back in a short cycle, while each intra-group antenna control amount changing slowly compared with the inter-group antenna control amount is updated and fed back in a longer cycle. In other words, since the change of inter-group diversity control with a low fading correlation is faster than that of the data speed of feedback information, the frequency of updates is made large. However, since the change of intra-group diversity control with a high fading correlation is slower than that of the data speed of feedback information, the frequency of updates is made small.

[0043] Since each intra-group antenna control amount has been related to the angle of the mobile station with respect to the base station, in a macro-cell system with a relatively large cell radius, the deviation of an input angle becomes negligibly small. Therefore, a specific intra-group antenna control amount can also be used as the intra-group antenna control amount of another group. Specifically, transmitting only the intra-group control information of one specific group and controlling the antennas in the other group using this information can further reduce an amount of feedback information.

[0044] Fig. 4 shows the configuration of one preferred embodiment of the present invention.

[0045] A case where the number of antennas N=4 and the number of antenna groups M=2 is described. Apilot signal generation unit 30 generates N=4 pilot signals $P_1(t)$, $P_2(t)$, $P_3(t)$ and $P_4(t)$, and each of the signals is transmitted from one of transmitting antennas 31. These pilot signals use mutually orthogonal bit sequences.

[0046] Each transmitting antenna 31 transmits the pilot signal to a mobile station. In the mobile station, a receiving antenna 32 receives the four pilot signals transmitted from each of four transmitting antennas, and a control amount calculation unit 33 estimates the channel of signals transmitted from each transmitting antenna 31 using the respective pilot signal. As a result, the channel impulse response vector is obtained from each signal and a weight vector that maximizes equation (5) is calculated. Since a method for calculating this weight vector is already publicly known, the description is omitted. When the weight vector is calculated, the control amount calculation unit 33 transfers the vector to a multiplex unit 34 as feedback information. The multiplex unit 34 multiplexes the feedback information with an upward data signal and transmits the information from a transmitting antenna 35. In a base station, a receiving antenna 36 receives the signal from the mobile station, and a feedback information extraction unit 37 extracts the feedback information from the signal. The extracted feedback information is inputted to an amplitude/phase control unit 38, each weight coefficient W_1, W_2 and w_3 included in the feedback information is multiplied to the respective downward transmitting data signal of each corresponding antenna, and the transmitting antennas 31 transmit the downward transmitting data signals. In this way, in this preferred embodiment, a closed loop for performing transmitting diversity control, including a base station and a mobile, is implemented.

[0047] Fig. 5 shows examples of a downlink pilot signal pattern in this preferred embodiment.

[0048] If each corresponding code is multiplied by each of the pilot signals P_1 through P_4 shown in Fig. 5, and the products of the entire pilot signal pattern are added up the result "0" is obtained. Specifically, the pilot signals P_1 through P_4 form a mutually orthogonal code word.

[0049] Each pilot signal's amplitude and phase change independently due to fading, and the combination of these signals is received by the antenna of a mobile station. A mobile-station receiver can calculate the channel response estimation values h_1, h_2, h_3 and h_4 of each pilot signal by correlating the incoming pilot signals with corresponding pilot signals $P_1(t)$, $P_2(t)$, $P_3(t)$ and $P_4(t)$, respectively, that are stored in advance on the mobile station side and by averaging the obtained correlations.

[0050] Fig. 6 shows both an example configuration of base-station transmitting antennas according to this preferred embodiment and antenna control information thereof.

[0051] Fig. 6A shows the transmitting antenna configuration of a base station. It is assumed that antennas ANT1 and ANT2 form group 1, and antennas ANT3 and ANT4 form group 2. It is also assumed that antennas ANT1 and ANT3 are the reference antenna of groups 1 and 2, respectively. It is further assumed that antenna ANT1 is also the reference antenna of all the groups 1 and 2. Antennas ANT1 and ANT2 are located apart from each other by one wavelength. Antennas ANT3 and ANT4 are also located apart from each other by one wavelength. Antennas ANT1 and ANT3 are located apart from each other by 20 wavelengths. Antennas ANT2 and ANT4 are also located apart from each other by 20 wavelengths.

[0052] Here, the spatial correlative characteristic of a base-station antenna is described.

[0053] If the input angles of signals from mobile stations are uniformly distributed with dispersion $\Delta\phi$, the envelope correlation coefficient of input waves is expressed as follows. In the equation, d represents the distance between two antennas.

$$\rho = \left(\frac{\sin X}{X} \right)$$

$$X = \frac{\pi d \Delta\phi}{\lambda}$$

[0054] The angle dispersion $\Delta\phi$ of each input signal observed at the base station in a macro-cell environment is approximately 3 degrees. Fig. 7 shows the envelope correlation coefficient in this case. It is seen from Fig. 7 that at $d \cong 19\lambda$ the input signals become uncorrelated. Therefore, according to the present invention, fading correlation can be made low by setting the distance between antenna groups to approximately 19 wavelengths or more. Fading correlation can also be made high by setting the distance between antennas in each group to one wavelength or less.

[0055] However, fading correlation is affected by a variety of factors, such as the height at which the antenna is installed, the size of the antenna and the like. Therefore, it is acceptable if the antennas are installed so that the distance between any two antennas in the same group is approximately the wavelength of an incoming signal. However, a person having ordinary skill in the art should set the distance between groups so that fading correlation is almost "0" in any situation.

[0056] Description will return to Fig. 6. In the following description it is assumed that amplitude is not controlled and only phase is controlled. Specifically, only a phase amount ϕ_i is controlled by assigning a_1=1 to $w_i = a_i e^{j\phi_i}$. As shown in Fig. 6B, each of the control amount ϕ_1 of antenna ANT2 using antenna ANT1 as a reference, the control amount ϕ_2 of antenna ANT4 using antenna ANT3 as a reference and the control amount ϕ_3 of antenna ANT3 using antenna ANT1 as a reference is quantized and is transmitted to the base station as feedback information. If each of the control amounts is quantized using one bit, for example, the setting is as follows.

$$\begin{aligned} &\frac{-\pi}{2} < \phi_i \le \frac{\pi}{2} \Rightarrow \phi_i^Q = 0 \\ &\frac{\pi}{2} < \phi_i \le \frac{3\pi}{2} \Rightarrow \phi_i^Q = \pi \end{aligned} \qquad (11)$$

[0057] In the expression, ϕ_i^Q is a quantized control amount.

[0058] Figs. 8 through 11 show examples of the transmission format of feedback information in this preferred embodiment.

[0059] It is assumed that if ϕ_i^Q=0, feedback information b_i=0 and that if $\phi_i^Q=\pi$, feedback information b_i=1. As shown in Fig. 8, this feedback information is multiplexed with an upward channel so that the transmission rate of b_3 may become higher than the transmission rate of b_1 or b_2 and is transmitted to a base station. One frame of length 10ms is composed of 15 slots in compliance with the W-CDMA frame format. This transmission format transmits feedback information of one bit in each slot. Format1 transmits both one b_1 and one b_2 in one frame, and format2 transmits both two b_1 and two b_2 in one frame.

[0060] In the base station, the phase control of each transmitting antenna is conducted using the feedback information received in an uplink channel. A corresponding antenna is directly controlled by the feedback information received in the immediately previous slot. In this case, antennas other than the corresponding antenna store the latest feedback information and use the information for their control.

[0061] However, ANT4 shown in Fig. 6A is controlled not only by control amount d2, but also by the control amount d3 of ANT 3. Specifically, ANT4 is frequently controlled by d3 and is also controlled by d2 less frequently. This description also applies to ANT4 shown in Fig. 6B.

[0062] Filtering feedback information can also reduce the number of transmission errors and the number of quantization errors. For example, for the filtering, a method using the average value of the control amount of the feedback information received in the immediately previous slot and the control amount of the feedback information received in receiving slots before the immediately previous slot is used.

[0063] As the feedback information of an intra-group antenna control amount, an updated control amount is transmitted every time the feedback information is transmitted. However, for example, alternatively, in the same frame, the same feedback information can also be repeatedly transmitted. In this case, the number of transmission errors in the base station can be reduced by combining a plurality of pieces of feedback information received in the frame.

[0064] Since each intra-group antenna control amount relates to the angle of a mobile station against the base station

in a macro-cell system with a to some extent large cell radius, the deviation of the input angle within a group is negligibly small. Therefore, if control is exercised within each group using the same intra-group antenna control amount, there is no problem. Therefore, transmitting only the intra-group control information of one specific group and controlling other groups using this information can further reduce an amount of feedback information.

[0065] Fig. 9 shows a feedback information transmission format used to transmit only b_1 as intra-group control information. Format3 transmits two b_1 in one frame and format 4 transmits four b_1 in one frame.

[0066] In this preferred embodiment too, as the feedback information of an intra-group antenna control amount, an updated control amount can be transmitted every time the feedback information is transmitted. Alternatively, for example, the same feedback information can be repeatedly transmitted within the same frame.

[0067] Another transmission format in which a control amount calculated in a mobile station is quantized using a plurality of bits is described below.

[0068] Fig. 10 shows the feedback information transmission format in which b_1 and b_3 are quantized using three bits and four bits, respectively. Tables 1 and 2 of Fig. 11 show the correspondence between the feedback information b_3 of an inter-group antenna control amount and a control amount. Table 3 shows the correspondence between the feedback information b_1 of an intra-group antenna control amount and a control amount.

[0069] In this example, only the feedback information of an intra-group antenna control amount b_1 is transmitted using the format shown in Fig. 9. As is clearly seen from Tables 1 and 2 of Fig. 11, feedback information bit b_3 is composed of four bits; three bits of $b_3(3)$ through $b_3(1)$ representing a phase control amount and one bit of $b_3(0)$ representing an amplitude control amount. Format5 shown in Fig. 10 includes feedback information bit b_3 in one frame. However, three words of feedback information bit b1 are composed of three bits of b_1 (2) through b_1 (0) representing a phase control amount. According to formats shown in Fig. 10, three bits of feedback information bit b_1 are distributed and located in one frame, and all the three bits together form one word.

[0070] Fig. 12 shows an example configuration of a mobile station that transmits feedback information to a base station according to the formats shown in Figs. 8 through 11.

[0071] On receipt of a signal from a base station via its receiving antenna, a mobile station branches the receiving signal into two signals and inputs one signal and the other signal to a data channel despreading unit 41 and a pilot channel despreading unit 44, respectively. The data channel dispreading unit 41 despreads the data channel signal and inputs the signal to both a channel estimation unit 42 and a receiver 43. The receiver 43 reproduces the downlink data signal, based on the channel estimation result of the channel estimation unit 42 and presents the signal to a user as voice or data. The pilot channel dispreading unit 44 despreads the incoming signal using a pilot channel dispreading code and inputs the signal to a channel estimation unit 45. The channel estimation unit 45 correlates the despread signal to each pilot signal pattern and obtains channel estimation values H=[h_1, h_2, h_3 and h_4] for paths from each transmitting antenna to the mobile station. A control amount calculation unit 46 calculates a weight vector based on these channel estimation values and determines feedback information to be transmitted. A multiplex unit 47 multiplexes this feedback information with an upward control channel. A data modulation unit 48 modulates the feedback information. A spreading modulation unit 49 spread-modulates the feedback information. Then, the feedback information is transmitted to the base station from a transmitting antenna 50.

[0072] In Fig. 13, the same reference numbers are attached to the same constituent components as those in Fig. 4 and their descriptions are omitted.

[0073] In this preferred embodiment, a base station uses both uplink feedback information and an uplink channel arriving method estimation result as intra-group antenna control information. In the base station, input direction estimation units 62 and 63 estimate the arriving direction of an incoming signal based on an uplink receiving signal received by an array antenna (a plurality of antennaslements used in transmission diversity: transmitting/receiving antenna 60). Since arriving direction strongly depends on the angle of a mobile station against a base station, a method for setting the direction of a downlink transmitting beam (direction in which the strength of a wave transmitted from an antenna is large) to this uplink signal input direction is known. However, in a system where uplink and downlink frequencies are different, this assumption does not always hold true and depends on the propagation environment.

[0074] Upon receipt of the uplink feedback information via an antenna 60, a receiving processing unit 61 performs the despeading and the like of the uplink feedback information and relays the information to a feedback information extraction unit 37. When the feedback information extraction unit 37 extracts a control amount from the uplink feedback information, an amplitude/phase control unit 38' compares the control amount with the arriving direction estimation value and determines to use either the control amount received from the uplink line or the arriving direction estimation value. Then, the unit 38' controls the amplitude/phase of a transmitting signal.

[0075] As shown in Fig. 14, in this preferred embodiment, if the intra-group phase difference is not within a specific range [θ-Δ,θ-Δ] with the arriving direction estimation result θ of the uplink channel as a center, control is exercised using only the arriving direction estimation result θ since the control amount by the upward feedback information is related to an uplink channel arriving direction estimation result θ. Specifically, if the control amount in the feedback information is too far from the arriving direction estimation result θ, it is judged that a bit error or the like has occurred

during transmission of the feedback information, and the feedback information is inaccurate. Then, the feedback information is discarded and only phase is controlled using the arriving direction estimation result θ.

[0076] Alternatively, a control amount in the uplink feedback information of intra-group phase difference information can be sampled for a prescribed time period. If it is judged that variance of the samples is large (for example, specifically, if the samples are dispersed more widely than a specific predetermined threshold value), control can be exercised using only the arriving method estimation result θ without utilizing the feedback information.

[0077] Fig. 15 shows the configuration of the third preferred embodiment of the present invention.

[0078] In Fig. 15, the same reference numbers are attached to the same components as those in Fig. 4, and their descriptions are omitted.

[0079] In this case, the transmitting powers of pilot signals P_1 and P_3 are set smaller than the transmitting powers of pilot signals P_2 and P_4, respectively. In this preferred embodiment, this is implemented by multiplying pilot signals P_2 and P_4 by a coefficient α ($0<\alpha<1$). Although pilot signals P_2 and P_4 are needed to estimate channel impulse response vectors $\underline{h}_2$ and $\underline{h}_4$, $\underline{h}_2$ and $\underline{h}_4$ have high fading correlations to $\underline{h}_1$ and $\underline{h}_3$, respectively. Therefore, $\underline{h}_2/\underline{h}_1$ and $\underline{h}_4/\underline{h}_3$ that are normalized by them strongly depend on an angle of the mobile station against the base station. Since these values fluctuate slowly compared with fading fluctuation, estimation accuracy can be improved by taking a long time average of pilot signals P_2 and P_4 even if incoming power on the mobile station side is low. Both ϕ_1 and ϕ_2 are calculated as follows.

$$\phi_1 = \underline{h}_2/\underline{h}_1,\ \phi_2 = \underline{h}_4/\underline{h}_3 \qquad (12)$$

[0080] Since interference to data signals by pilot signals can be suppressed to a low level by setting the transmitting powers of pilot signals P_2 and P_4 to a low level, transmission capacity can be increased.

[0081] Since both $\underline{h}_2/\underline{h}_1$ and $\underline{h}_4/\underline{h}_3$ depend on the angle of the mobile station against the base station and fluctuate more slowly than a fading fluctuation, estimation accuracy can be improved by taking a long time average of pilot signals P_2 and P_4 even if an incoming power is low. For example, estimation values $\phi_1(n)$, $\phi_2(n)$ and ϕ_3 (n) at the n-th slot can be calculated as follows. In these equations, N is the estimated average number of slots of estimation values ϕ_1 (n) and $\phi_2(n)$.

$$\phi_1(n) = \frac{1}{N}\sum_{i=0}^{N-1}\frac{\underline{h}_2(n-i)}{\underline{h}_1(n-i)}$$

$$\phi_2(n) = \frac{1}{N}\sum_{i=0}^{N-1}\frac{\underline{h}_4(n-i)}{\underline{h}_3(n-i)}$$

$$\phi_3(n) = \frac{\underline{h}_3(n)}{\underline{h}_1(n)}$$

[0082] In this way, when both ϕ_1 and ϕ_2 are calculated by taking a N-times time (number of slots) average of ϕ_3, the same estimation accuracy as that of ϕ_3 can be obtained even if α = 1/N. Specifically, in case N=4, α = 1/4 can be assigned.

Industrial Applicability

[0083] If the number of transmitting antennas is increased by utilizing differences in the fluctuation rate of control information, the following effects can be obtained.

- The increase in the amount of upward feedback information can be suppressed.
- Characteristics degrade little in the case of a high fading frequency.
- The antenna installation space of a base station can be reduced.

Claims

1. A transmitting diversity communications apparatus, including a base station adopting a transmitting diversity method, for controlling transmitting signals according to information from a mobile station, comprising:

   antenna means composed of a plurality of antenna groups, each group consisting of a plurality of antennas located close to one another so that fading correlation between the antennas is high, and the antenna groups are located apart from one another so that fading correlation between the groups is low; and
   control means for receiving both first control information for intra-group antenna control, with a low transfer rate and second control information for inter-antenna group control, with a high transfer rate that are transmitted from a mobile station, and controlling a phase of a signal transmitted by the antenna means.

2. The transmitting diversity communications apparatus according to claim 1, wherein the mobile station determines a control amount of the phase using pilot signals transmitted from the base station.

3. The transmitting diversity communications apparatus according to claim 1, wherein said control means also controls amplitude in addition to the phase.

4. The transmitting diversity communications apparatus according to claim 3, wherein the mobile station determines control amounts of both the phase and amplitude using pilot signals transmitted from the base station.

5. The transmitting diversity communications apparatus according to claim 4, wherein the mobile station estimates a channel response from each antenna to the mobile station by correlating a pilot signal from the base station to a known pilot signal on a mobile station side and calculating the control amount using this channel response estimation value.

6. The transmitting diversity communications apparatus according to claim 1, wherein the mobile station transmits information describing the difference in channel response estimation values between each intra-group antenna of said antenna means and a reference antenna and information describing the difference in channel response estimation values between each antenna group and a reference antenna of a specific antenna group to the base station as the first and second control information, respectively.

7. The transmitting diversity communications apparatus according to claim 1, wherein the mobile station transmits control information about each antenna group and control information about an intra-group antenna within a specific antenna group to the base station as the second and first control information, respectively.

8. The transmitting diversity communications apparatus according to claim 1, wherein said control means controls the transmitting of a signal from the base station using an input direction estimation result of an uplink channel signal in addition to the first and second control information.

9. The transmitting diversity communications apparatus according to claim 8, wherein if transmitting signal control amounts obtained from the first and second control information do not fall within a specific range with an arriving direction estimation result of the uplink channel signal as a center, transmission is controlled using the input direction estimation result.

10. The transmitting diversity communications apparatus according to claim 8, wherein if transmitting signal control amount dispersion obtained from the first control information is larger than a prescribed value, transmission is controlled using only an arriving direction estimation result.

11. The transmitting diversity communications apparatus according to claim 1, wherein control is exercised by a filtering result using both currently received first and second information and one or more previously received first and second control information.

12. The transmitting diversity communications apparatus according to claim 1, wherein the power of a signal transmitted from an antenna other than a reference antenna is set at a lower level than the power of a signal transmitted from a reference antenna of each antenna group.

13. A transmitting diversity communications apparatus, including a base station adopting a transmitting diversity meth-

od, for controlling transmitting signals according to information from a mobile station, comprising:

antenna means composed of a plurality of antenna groups, each group consisting of a plurality of antennas located close to one another so that fading correlation between the antennas is high, and the antenna groups are located apart from one another so that fading correlation between the groups is low; and
control means for receiving both first control information for intra-group antenna control, with a low transfer rate and second control information for inter-antenna group control, with a high transfer rate that are transmitted from a mobile station, and controlling both amplitude and phase of a signal transmitted by the antenna means.

14. A mobile station of a transmitting diversity communications apparatus, including a base station adopting a transmitting diversity method, for controlling transmitting signals according to information from a mobile station, comprising:

receiving means for receiving a signal transmitted from an antenna means composed of a plurality of antenna groups, each group consisting of a plurality of antennas located close to one another so that fading correlation between the antennas is high, and the antenna groups are located apart from one another so that fading correlation between the groups is low;
antenna specifying means for identifying an antenna that has transmitted the received signal; and
transmitting means for transmitting first control information about intra-group antenna control of the received signal to the base station at a prescribed transfer rate and transmitting second control information about inter-group antenna control of the received signal to the base station at a higher transfer rate than the prescribed transfer rate.

15. A transmitting diversity communications method, including a base station adopting a transmitting diversity method, for controlling transmitting signals according to information from a mobile station, comprising:

providing a plurality of antenna groups, each group consisting of a plurality of antennas, placing the antennas in the same group close to one another so that fading correlation between the antennas in the same group is high and placing the antenna groups apart from one another so that fading correlation between the groups is low; and
receiving both first control information for intra-group antenna control, with a low transfer rate and second control information for inter-antenna group control, with a high transfer rate that are transmitted from a mobile station and controlling the phase of a signal transmitted by the antenna unit.

16. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
some of the plurality of antennas of the base station have a location relation in which, with respect to one antenna, other antennas are placed where fading correlation is high, and where fading correlation is low, and
the mobile station transmits phase control information about an antenna located in the position having a high fading correlation and phase control information about an antenna located in the position having a low fading correlation to the base station with low frequency and high frequency, respectively.

17. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
some of the plurality of antennas of the base station have a location relation in which, with respect to one antenna, other antennas are placed where fading correlation is high, and where fading correlation is low, and
the mobile station transmits phase control information about an antenna located in the position having a high fading correlation to the base station with lower frequency than frequency of phase control information of an antenna located in the position having a low fading correlation.

18. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
all the plurality of antennas except a specific antenna of the base station are located in positions having a specific fading correlation to the antenna, and
the mobile station transmits phase control information about the antennas except the specific antenna to the base station with frequency corresponding to the specific fading correlation.

19. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of

antennas of a base station according to phase control information from a mobile station, wherein

all the plurality of antennas except a specific antenna of the base station are located in positions having a high fading correlation to the antenna, and

the mobile station transmits phase control information about the antennas except the specific antenna to the base station with low frequency.

20. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein

all the plurality of antennas except a specific antenna of the base station are located in positions having a low fading correlation to the antenna, and

the mobile station transmits phase control information about all the antennas except the specific antenna to the base station with high frequency.

21. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where some of the plurality of antennas and the other antennas except a specific antenna are located in positions having a high fading correlation and in positions having a low fading correlation, respectively, to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting phase control information about an antenna located in a position having a high fading correlation and phase control information of an antenna located in a position having a low fading correlation to the base station with low frequency and high frequency, respectively.

22. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where some of the plurality of antennas and the other antennas except a specific antenna are located in positions having a high fading correlation and in positions having a low fading correlation, respectively, to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting phase control information about an antenna located in a position having a high fading correlation to the base station with lower frequency than frequency of phase control information of an antenna located in a position having a low fading correlation.

23. A mobile station of a communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas on a base station side where all the plurality of antennas except a specific antenna are located in positions having a specific fading correlation to the antenna according to phase control information from a mobile station, comprising:

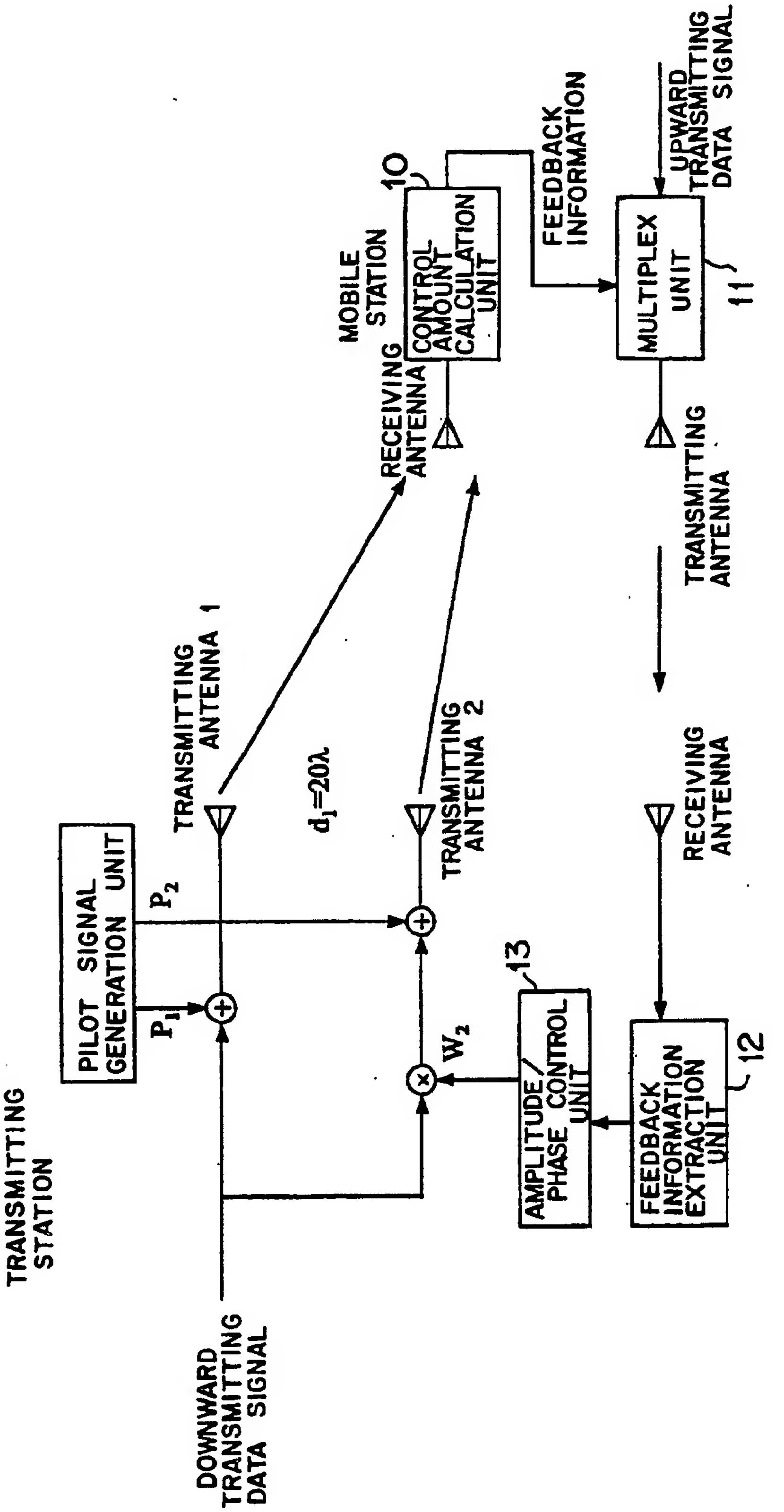
control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting the corresponding antenna phase control information to the base station with frequency corresponding to the fading correlation.

24. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where the other of the plurality of antennas are located in positions having a high fading correlation with one antenna according to phase control information from a mobile station, comprising:

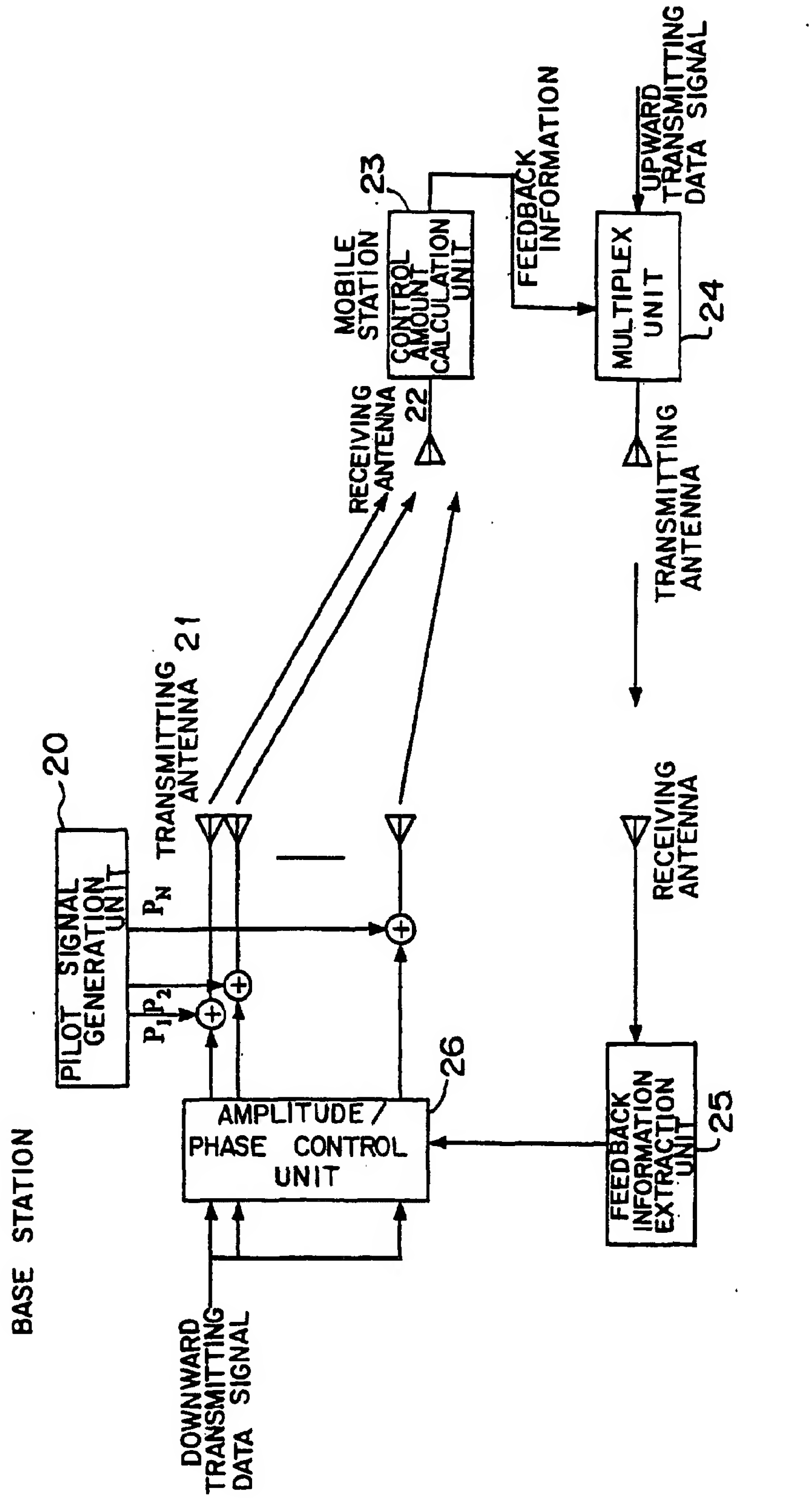
control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting antenna phase control information about all the antennas except the specific antenna to the base station with low frequency.

25. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where all the plurality of antennas except a specific antenna are located in positions having a low fading correlation to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting phase control information about all the antennas except a specific antenna to the base station with high frequency.


TRANSMITTING STATION
PILOT SIGNAL GENERATION UNIT
P_1
P_2
TRANSMITTING ANTENNA 1
$d_1=20\lambda$
TRANSMITTING ANTENNA 2
W_2
13
AMPLITUDE/ PHASE CONTROL UNIT
FEEDBACK INFORMATION EXTRACTION UNIT
12
RECEIVING ANTENNA
DOWNWARD TRANSMITTING DATA SIGNAL
MOBILE STATION
RECEIVING ANTENNA
CONTROL AMOUNT CALCULATION UNIT
10
FEEDBACK INFORMATION
MULTIPLEX UNIT
11
UPWARD TRANSMITTING DATA SIGNAL
TRANSMITTING ANTENNA

F I G. 1


BASE STATION
DOWNWARD TRANSMITTING DATA SIGNAL
AMPLITUDE / PHASE CONTROL UNIT
26
PILOT SIGNAL GENERATION UNIT
20
P_1 P_2
P_N
TRANSMITTING ANTENNA
21
RECEIVING ANTENNA
FEEDBACK INFORMATION EXTRACTION UNIT
25
MOBILE STATION
RECEIVING ANTENNA
22
CONTROL AMOUNT CALCULATION UNIT
23
FEEDBACK INFORMATION
MULTIPLEX UNIT
24
UPWARD TRANSMITTING DATA SIGNAL
TRANSMITTING ANTENNA

F I G. 2

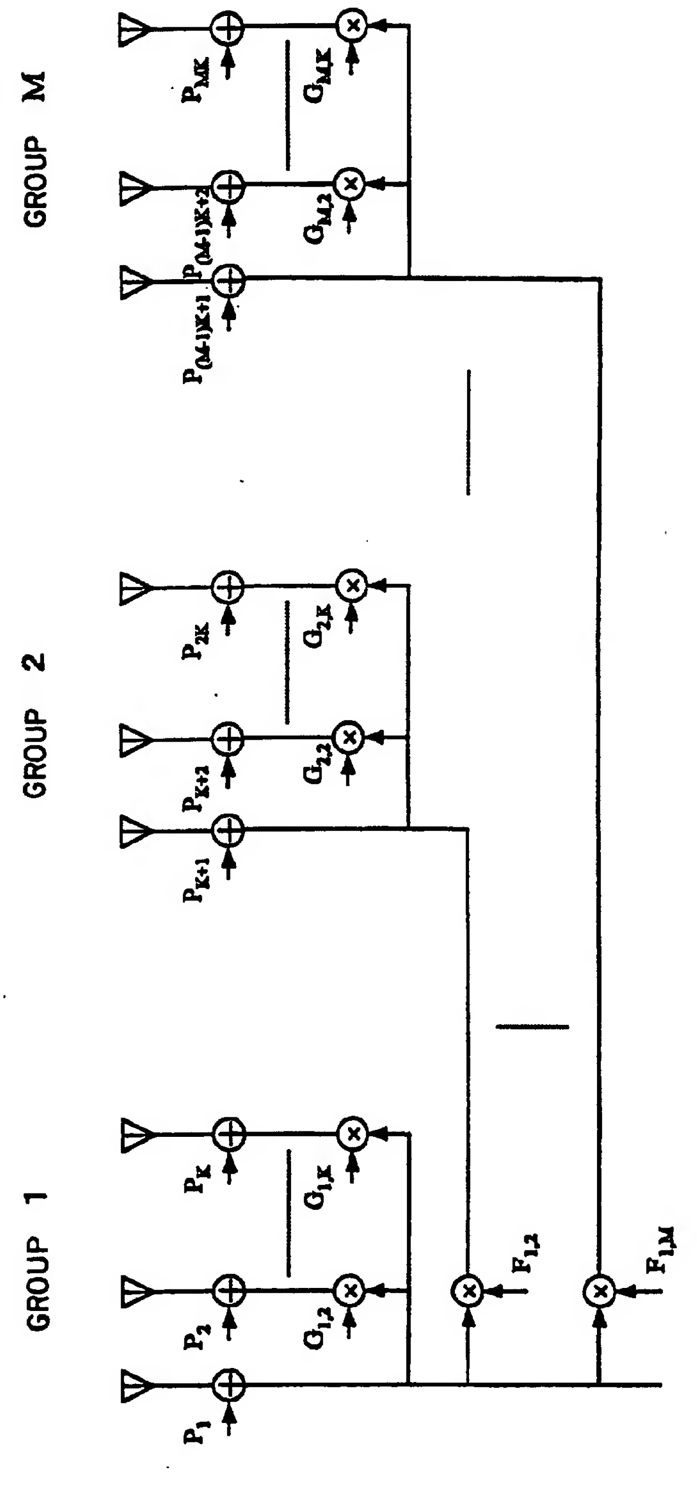

GROUP 1
GROUP 2
GROUP M
P_1
P_2
P_K
P_{K+1}
P_{K+2}
P_{2K}
$P_{(M-1)K+1}$
$P_{(M-1)K+2}$
P_{MK}
$G_{1,2}$
$G_{1,K}$
$G_{2,2}$
$G_{2,K}$
$G_{M,2}$
$G_{M,K}$
$F_{1,2}$
$F_{1,M}$

FIG. 3

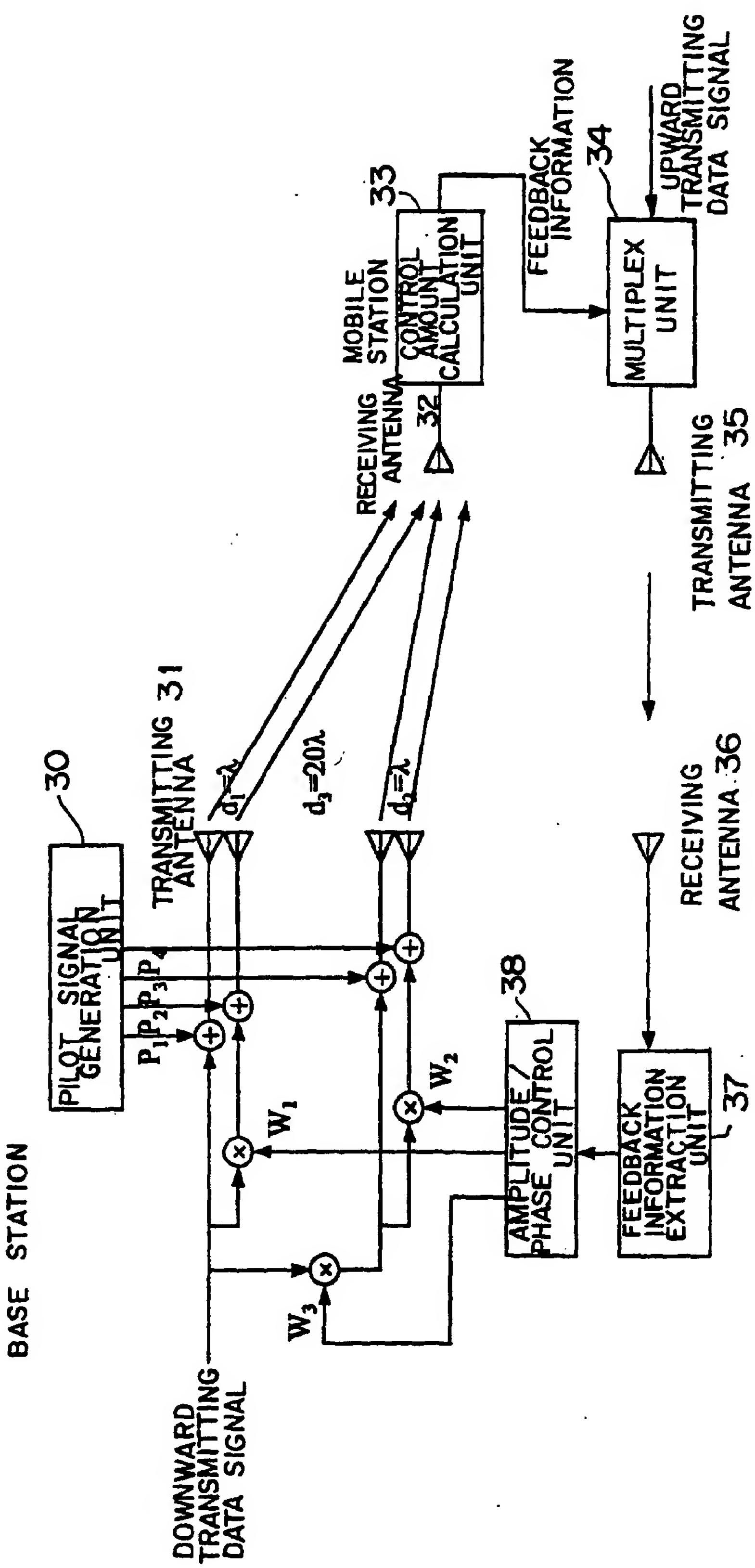
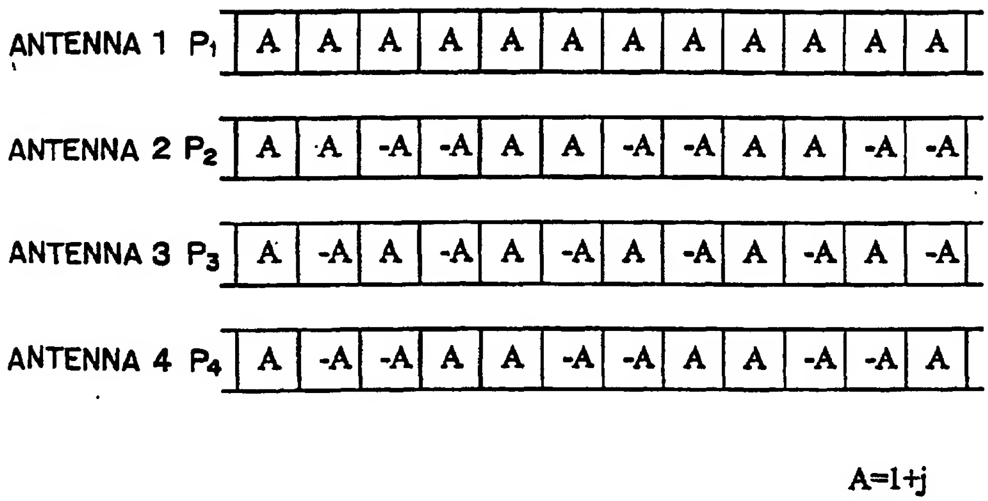
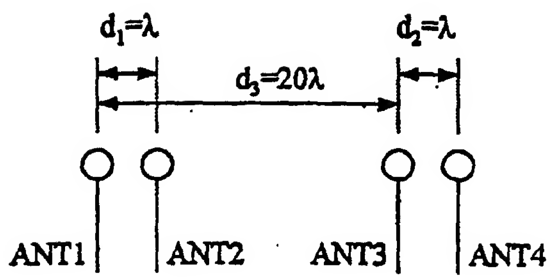

BASE STATION
DOWNWARD TRANSMITTING DATA SIGNAL
PILOT SIGNAL GENERATION UNIT
30
P_1 P_2 P_3 P_4
TRANSMITTING ANTENNA 31
$d_1=\lambda$
$d_3=20\lambda$
$d_2=\lambda$
W_1
W_2
W_3
AMPLITUDE/ PHASE CONTROL UNIT
38
FEEDBACK INFORMATION EXTRACTION UNIT
37
RECEIVING ANTENNA 36
MOBILE STATION
RECEIVING ANTENNA
32
CONTROL AMOUNT CALCULATION UNIT
33
FEEDBACK INFORMATION
MULTIPLEX UNIT
34
UPWARD TRANSMITTING DATA SIGNAL
TRANSMITTING ANTENNA 35
TRANSMITTING ANTENNA

FIG. 4

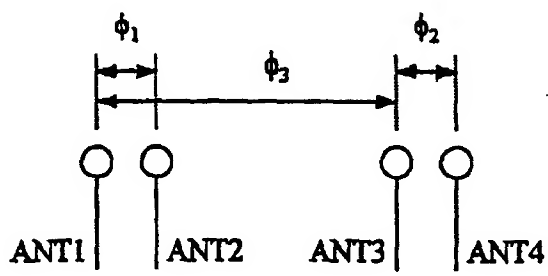

ANTENNA 1 P_1
A A A A A A A A A A A A
ANTENNA 2 P_2
A A -A -A A A -A -A A A -A -A
ANTENNA 3 P_3
A -A A -A A -A A -A A -A A -A
ANTENNA 4 P_4
A -A -A A A -A -A A A -A -A A
A=1+j

F I G. 5



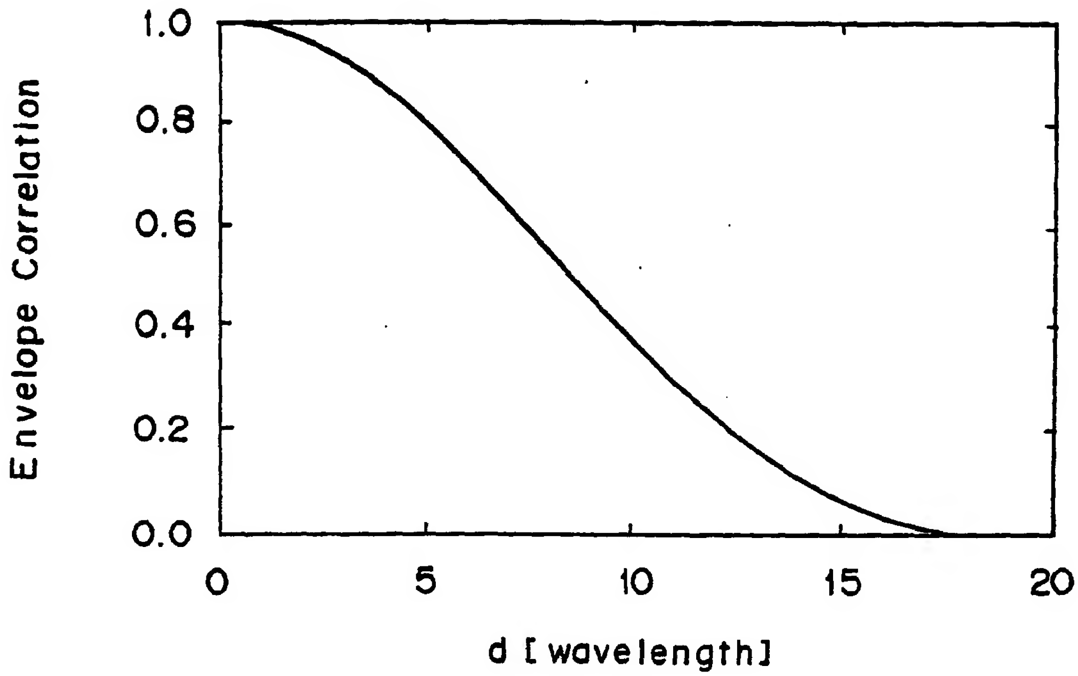
$d_1=\lambda$
$d_2=\lambda$
$d_3=20\lambda$
ANT1
ANT2
ANT3
ANT4

F I G. 6 A



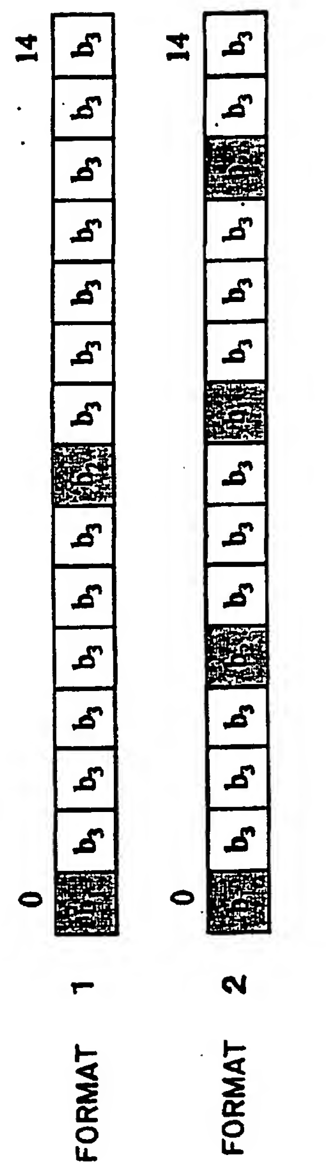
ϕ_1
ϕ_2
ϕ_3
ANT1
ANT2
ANT3
ANT4

F I G. 6 B

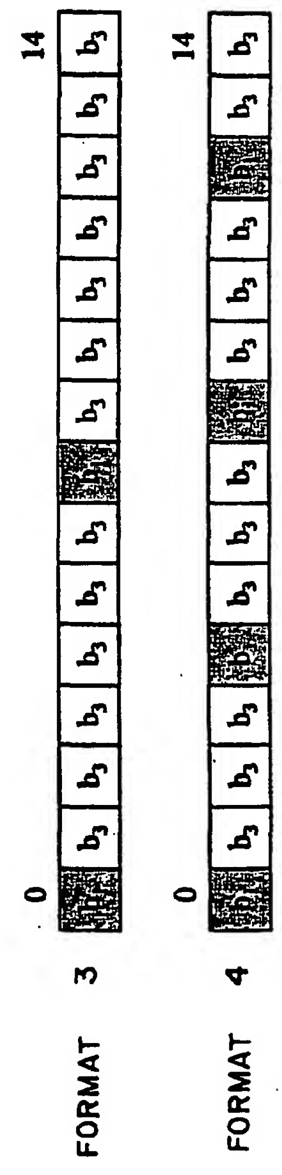


1.0
0.8
0.6
0.4
0.2
0.0
Envelope Correlation
0
5
10
15
20
d [wavelength]

F I G. 7



F I G. 8



F I G. 9

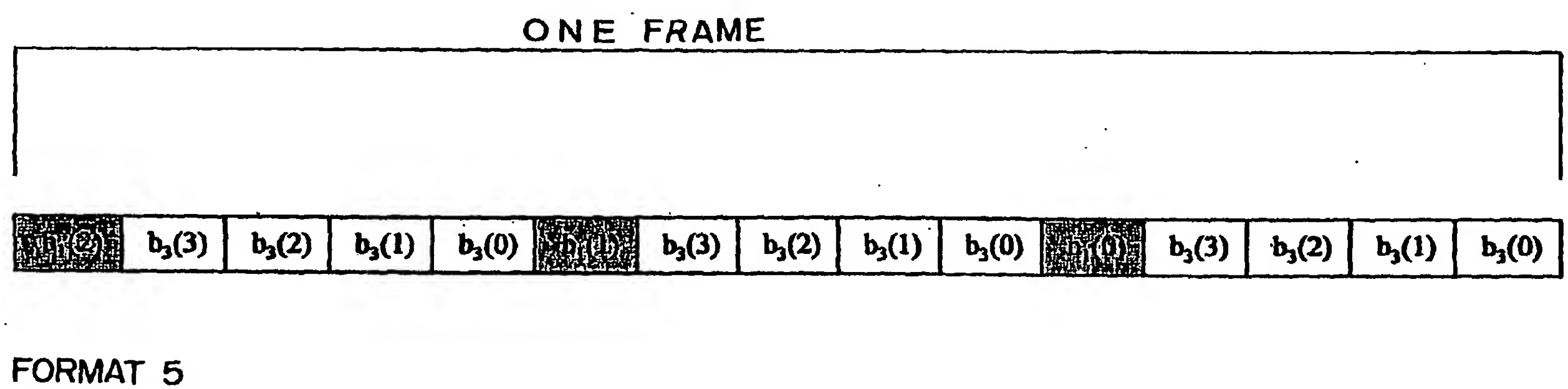

ONE FRAME
$b_3(3)$
$b_3(2)$
$b_3(1)$
$b_3(0)$
$b_3(3)$
$b_3(2)$
$b_3(1)$
$b_3(0)$
$b_3(3)$
$b_3(2)$
$b_3(1)$
$b_3(0)$
FORMAT 5

FIG. 10

TABLE 1. FEEDBACK BIT ($b_3(0)$)

$b_3(0)$	ANTENNA 1 AMPLITUDE	ANTENNA 2 AMPLITUDE
0	0.2	0.8
1	0.8	0.2

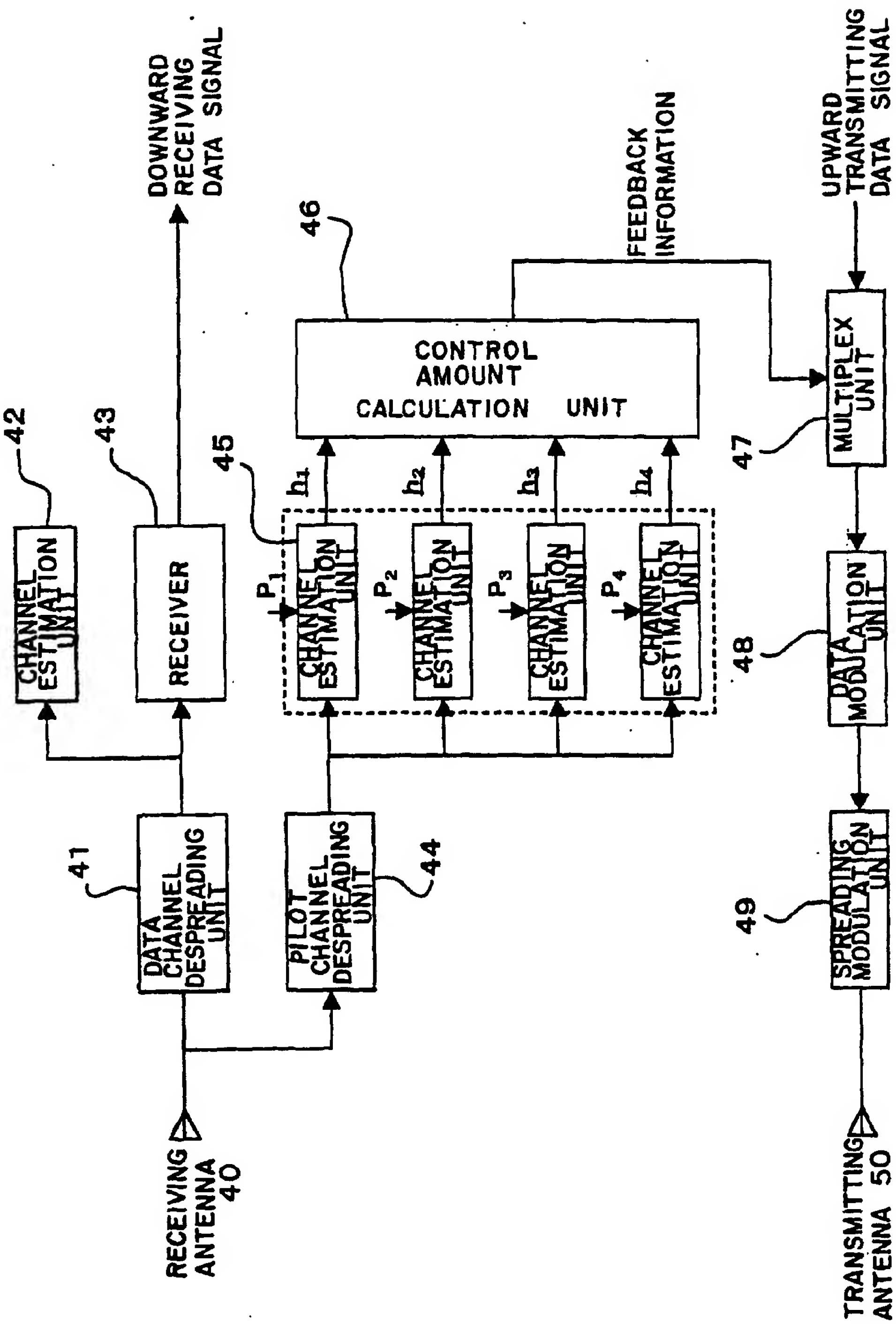
TABLE 2. FEEDBACK BIT ($b_3(3)$, $b_3(2)$, $b_3(1)$)

$b_3(3)$, $b_3(2)$, $b_3(1)$	INTER-ANTENNA PHASE DIFFERENCE (DEGREE)
000	180
001	-135
010	-90
011	-45
100	0
101	45
110	90
111	135

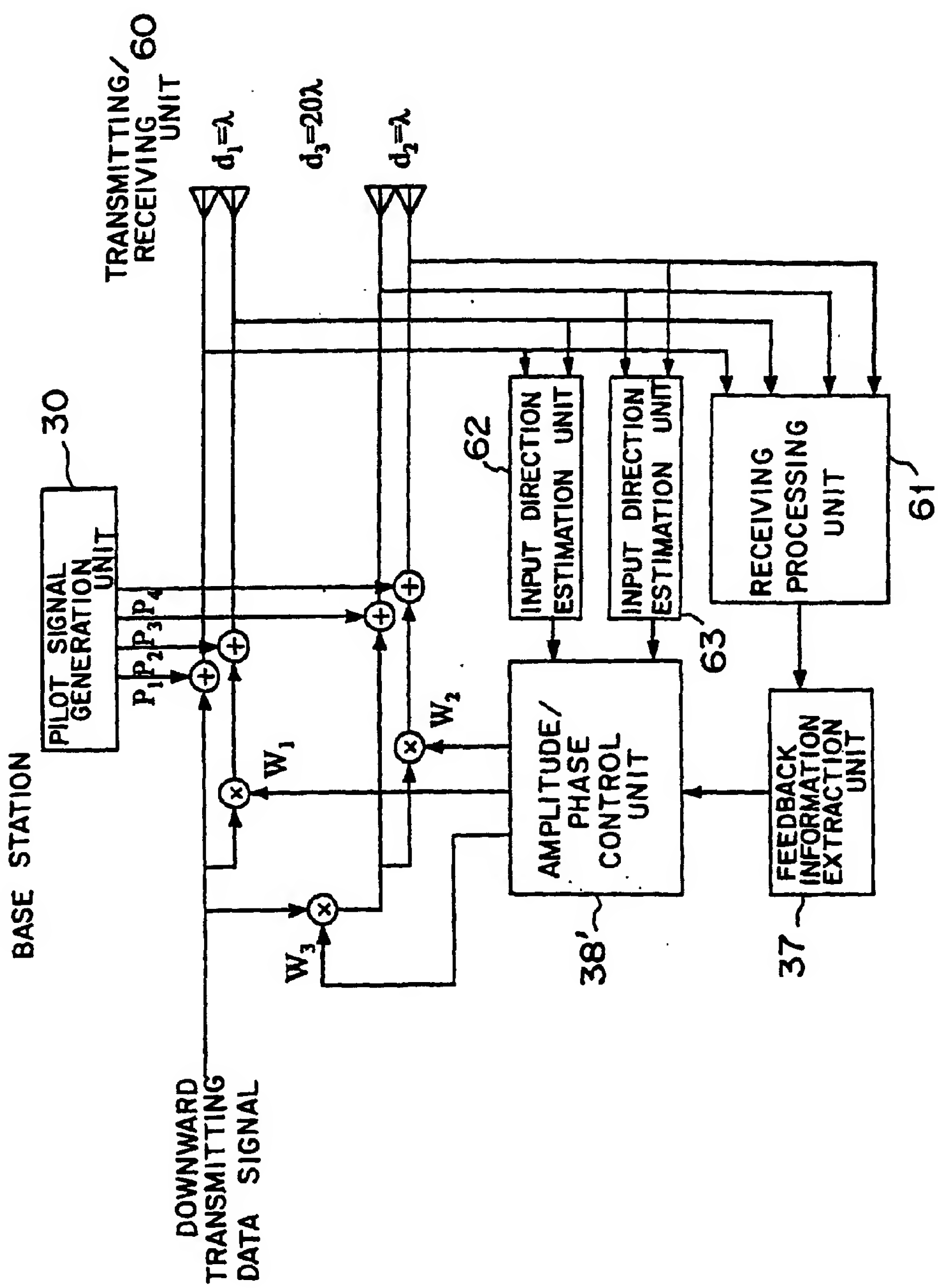
TABLE 3. FEEDBACK BIT ($b_1(2)$, $b_1(1)$, $b_1(0)$)

$b_1(2)$, $b_1(1)$, $b_1(0)$	INTER-ANTENNA PHASE DIFFERENCE (DEGREE)
000	180
001	-135
010	-90
011	-45
100	0
101	45
110	90
111	135

FIG. 11


RECEIVING ANTENNA 40
41
DATA CHANNEL DESPREADING UNIT
PILOT CHANNEL DESPREADING UNIT
44
42
CHANNEL ESTIMATION UNIT
43
RECEIVER
DOWNWARD RECEIVING DATA SIGNAL
45
P1
P2
P3
P4
CHANNEL ESTIMATION UNIT
CHANNEL ESTIMATION UNIT
CHANNEL ESTIMATION UNIT
CHANNEL ESTIMATION UNIT
h1
h2
h3
h4
46
CONTROL AMOUNT CALCULATION UNIT
FEEDBACK INFORMATION
UPWARD TRANSMITTING DATA SIGNAL
47
MULTIPLEX UNIT
48
DATA MODULATION UNIT
49
SPREADING MODULATION UNIT
TRANSMITTING ANTENNA 50

F I G. 12


BASE STATION
PILOT SIGNAL GENERATION UNIT
30
P_1 P_2 P_3 P_4
TRANSMITTING/ RECEIVING UNIT
60
$d_1=\lambda$
$d_3=20\lambda$
$d_2=\lambda$
W_1
W_2
W_3
INPUT DIRECTION ESTIMATION UNIT
62
INPUT DIRECTION ESTIMATION UNIT
63
RECEIVING PROCESSING UNIT
61
AMPLITUDE/ PHASE CONTROL UNIT
38'
FEEDBACK INFORMATION EXTRACTION UNIT
37
DOWNWARD TRANSMITTING DATA SIGNAL

F I G. 13

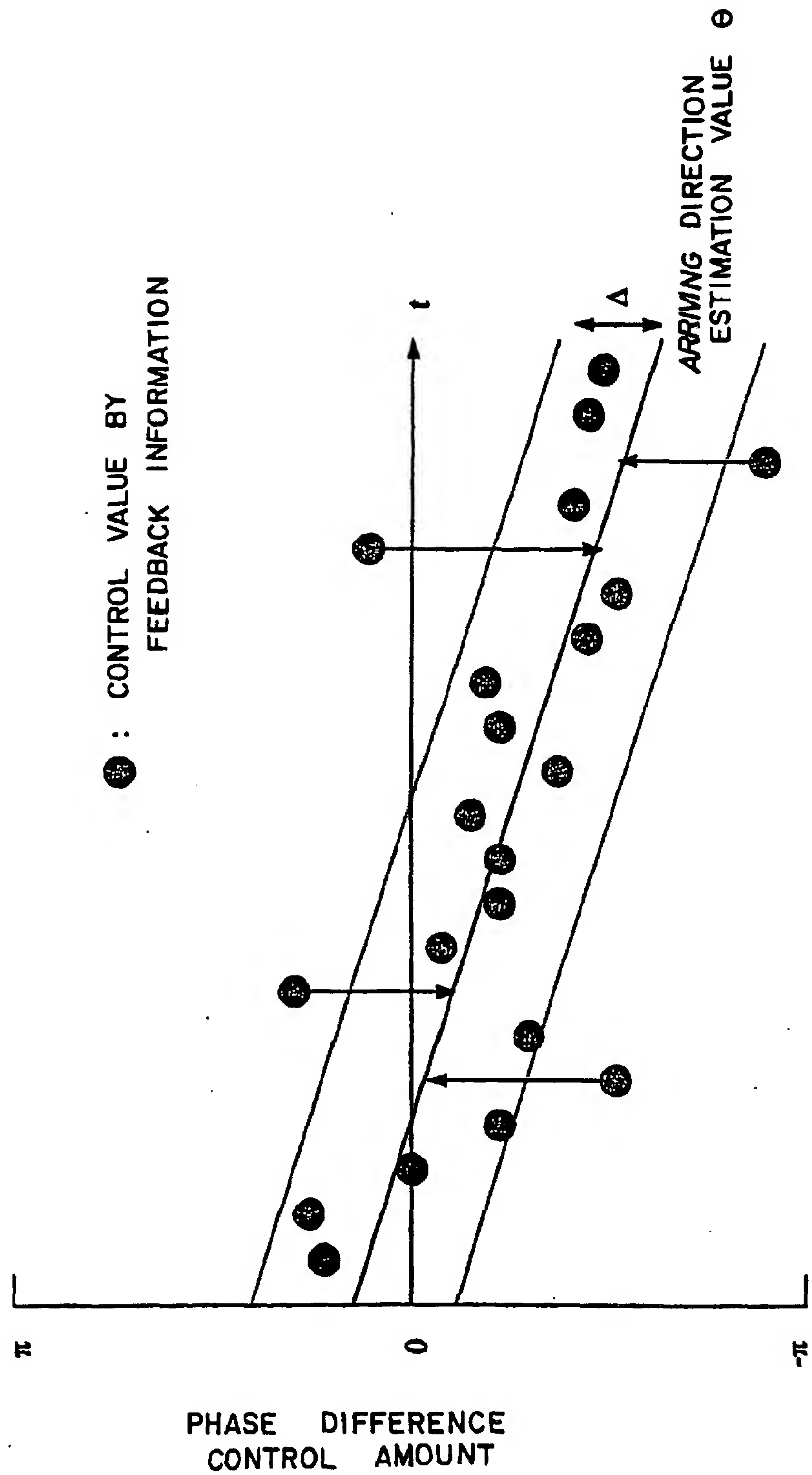

: CONTROL VALUE BY
FEEDBACK INFORMATION
t
Δ
ARRIVING DIRECTION
ESTIMATION VALUE θ
π
0
-π
PHASE DIFFERENCE
CONTROL AMOUNT

FIG. 14

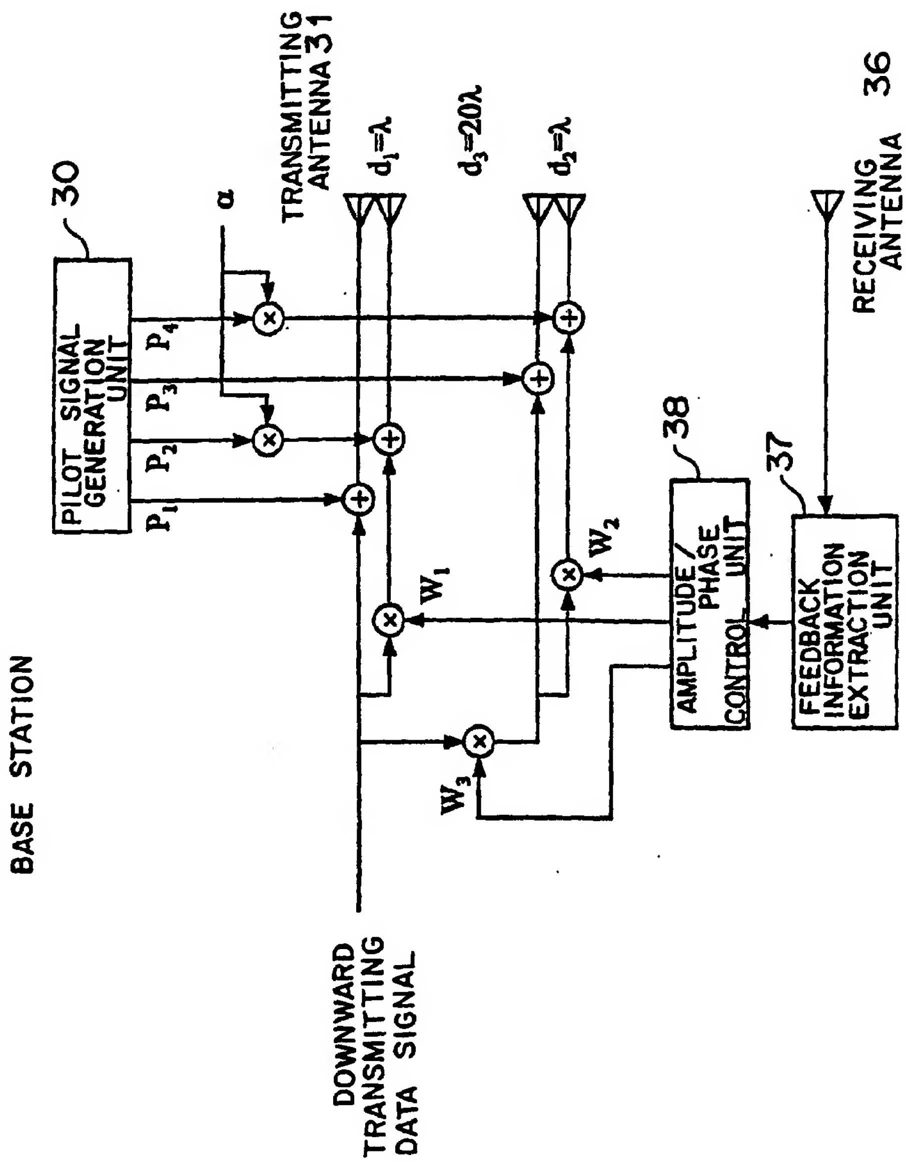

BASE STATION
30
PILOT SIGNAL GENERATION UNIT
P1
P2
P3
P4
α
TRANSMITTING ANTENNA 31
d1=λ
d3=20λ
d2=λ
RECEIVING ANTENNA
36
W1
W2
W3
38
AMPLITUDE/PHASE CONTROL UNIT
37
FEEDBACK INFORMATION EXTRACTION UNIT
DOWNWARD TRANSMITTING DATA SIGNAL

FIG. 15

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP00/05380

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl[7] H04B 7/06, 7/10, 7/26, H01Q 3/24

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl[7] H01Q 3/00- 3/46, 21/00-25/04
H04B 7/00, 7/02-7/12, 7/24-7/26, 113
H04L 1/02- 1/06, H04Q7/00-7/04

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
Jitsuyo Shinan Koho 1922-1996 Toroku Jitsuyo Shinan Koho 1994-2000
Kokai Jitsuyo Shinan Koho 1971-2000 Jitsuyo Shinan Toroku Koho 1996-2000

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y A	JP 58-87928 A (Nippon Telegr. & Teleph. Corp. <NTT>), 25 May, 1983 (25.05.83) (Family: none)	20,25 1-19,21-24
A	JP 10-190537 A (NEC Corporation), 21 July, 1998 (21.07.98) (Family: none)	1-25
A	JP 9-200115 A (Toshiba Corporation), 31 July, 1997 (31.07.97) (Family: none)	1-25

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
23 October, 2000 (23.10.00)	31 October, 2000 (31.10.00)

Name and mailing address of the ISA/	Authorized officer
Japanese Patent Office	
Facsimile No.	Telephone No.

Form PCT/ISA/210 (second sheet) (July 1992)

(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) EP 1 379 020 A1

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(43) Date of publication:
07.01.2004 Bulletin 2004/02

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AL LT LV MK RO SI

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  Serangoon Central Singapore 550208 (SG)

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80336 München (DE)

(54) A wireless communication apparatus and method

(57) A method and apparatus for achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of: providing a signal to be transmitted; space-time encoding the signal to produce at least two separate signals, each on a respective output; feeding each output signal to a multiple access transmit processor to produce an output signal; applying respective selected transmit beamforming weights to each output signal; feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal for transmission; feeding the summed signal to each of the multiple transmit antennae for transmission; transmitting the signals over respective physical channels; receiving the transmitted signal at at least a single receive antenna; feeding the transmitted signal to a multiple access receive processor to produce an output signal; and space-time decoding the received signal.

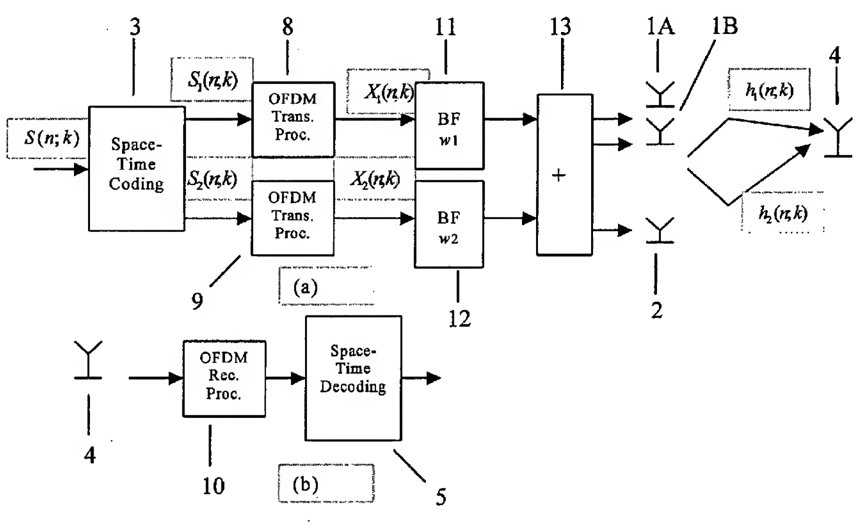

Figure5

EP 1 379 020 A1

Description

BACKGROUND OF THE INVENTION

[0001] The present invention relates in general to wireless communication systems and, more particularly, to improving the downlink performance of wireless communication systems.

[0002] Wireless mobile communications suffer from four major impairments: path loss, multipath fading, inter-symbol interference (ISI) and co-channel interference. Adaptive antennas can be used to suppress the effects of these factors to improve the performance of wireless communication systems. There are two types of adaptive antennas: diversity antennas and beamforming antennas. In a diversity antenna system, multiple low-correlation or independent fading channels are acquired in order to compensate multipath fading, thus achieving diversity gain. Beamforming antennas, on the other hand, provide beamforming gain by making use of spatial directivity, thus compensating for path loss to a certain extent and suppressing co-channel interference.

[0003] In a diversity antenna system, the antenna spacing is usually required to be large enough, e.g., 10λ in order to obtain low-correlation/independent fading channels, especially for small angular spread environments. However, beamforming antennas need to achieve spatial directivity, so the signals received at and/or transmitted from all antennas must be correlated. This means that for beamforming antenna, the antenna spacing should usually be small, e.g. half wavelength for a uniform linear array (ULA). Because of the conflict between the required antenna spacings for diversity antenna systems and beamforming systems, a prejudice exists that diversity gain and beamforming gain cannot be achieved simultaneously.

SUMMARY OF THE INVENTION

[0004] It is an object of the present invention to seek to provide a wireless communication system benefiting simultaneously from both diversity gain and beamforming gain.

[0005] Accordingly, one aspect of the present invention provides a method of achieving transmit diversity gain in a communication system having a base station with multiple transmit antennae and a mobile terminal with a single receive antenna, the method comprising the steps of: providing a signal to be transmitted $s(n)$; space-time encoding the signal $s(n)$ to produce at least two separate signals $s_1(n),s_2(n)$, each on a respective output; feeding each output signal $s_1(n),s_2(n)$ to a zero-forcing pre-equaliser having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$; feeding the output signal $x_1(n)$, $x_2(n)$ of each pre-equaliser to a transmit antenna; transmitting the output signals $x_1(n)$, $x_2(n)$ over respective physical channels $h_1(k)$, $h_2(k)$; receiving the output signals $x_1(n)$, $x_2(n)$ at a single receive antenna; and space-time decoding the received signals, wherein the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(k)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

[0006] Preferably, the communications system is a time-division duplex system and the method includes the further step of deriving the real channel coefficients from uplink channel coefficients for use in selecting the functions $g_1(k)$, $g_2(k)$ of the pre-equalisers.

[0007] Conveniently, the step of deriving the real channel coefficients from uplink channel coefficients uses training symbols from the uplink channel.

[0008] Advantageously, the step of deriving the real channel coefficients from uplink channel coefficients uses blind techniques.

[0009] Preferably, the communications system is a frequency-division duplex system and the method includes the further step of deriving the real channel coefficients by sending a set of training symbols to the receive antenna of the mobile terminal, the mobile terminal estimating the real channel coefficients and feeding back channel coefficient information to the base station.

[0010] Another aspect of the present invention provides a base station with multiple transmit antennae for communicating with a mobile terminal having a single receive antenna over physical channels $h_1(k)$, $h_2(k)$, the base station comprising:

> a space-time encoder having an input of a signal to be transmitted $s(n)$ and at least two outputs each producing a separate signal $s_1(n),s_2(n)$; at least two zero-forcing pre-equalisers, each fed by a respective output signal $s_1(n),s_2(n)$ and having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$; and at least two transmit antennae, each being fed by the output signal $x_1(n)$, $x_2(n)$ of a respective one of the pre-equalisers, wherein the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(k)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

[0011] Preferably, the mobile terminal has a single receive antenna and a space-time decoder to decode the signals

received from the base station.

[0012] A further aspect of the present invention provides a method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with a single receive antenna, the method comprising the steps of: providing a signal to be transmitted $S(n;k)$; space-time encoding the signal $S(n;k)$ to produce at least two separate signals $S_1(n;k), S_2(n;k)$, each on a respective output; feeding each output signal $S_1(n;k), S_2(n;k)$ to a transmit processor to produce an output signal $X_1(n;k)$, $X_2(n;k)$; applying respective selected transmit beamforming weights to each output signal $X_1(n;k)$, $X_2(n;k)$; feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal X(n;k) for transmission; feeding the summed signal X(n;k) to each of the multiple transmit antennae for transmission; transmitting the signals X(n;k) over respective the physical channel h(n;k); receiving the received signal Y(n;k) at a single receive antenna; feeding the received signal Y(n;k) to a receive processor to produce an output signal; and space-time decoding the received signal.

[0013] Preferably, the respective transmit beamforming weights are selected as the eigenvectors corresponding to the two largest eigenvalues of the downlink channel covariance matrix (DCCM) of the physical channel h(n;k).

[0014] Conveniently, the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, and the transmit processors do not add cyclic prefixes and one of the output signals from the transmit processors is delayed by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto, the beamforming weights being chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

[0015] Advantageously, the physical channel h(n;k) consists of two time-delayed clustered rays, $h_1(n;k)$ and $h_2(n;k)$, the transmit processors have a cyclic prefix length of $\Delta\psi$ and one of the output signals from the transmit processors is delayed by ψ before the respective selected transmit beamforming weight is applied thereto, the beamforming weights being chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

[0016] Preferably, the method comprises the further steps of: estimating a power-delay-DOA profile for the channel h(n;k); and, based on the profile: determining the cyclic prefix length, $\Delta\psi$, to be added by the transmit processors; determining the delay ψ; and determining the transmit beamforming weights.

[0017] Advantageously, the method comprises the further step of estimating the downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

[0018] Conveniently, the method comprises the further steps of: estimating the downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights; estimating a power-delay-DOA profile for channel h(n;k); and, based on the profile: determining the length, $\Delta\psi$, of the cyclic prefix to be added by the transmit processors; determining the delay ψ; and determining the transmit beamforming weights.

[0019] A further aspect of the present invention provides a base station with multiple transmit antennae for communicating with a mobile terminal having a single receive antenna over physical channel h(n;k) having two time-delayed rays, h_1(n;k) and h_2(n;k), the base station comprising:

> a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal; at least two transmit processors each receiving one of the outputs from a respective space-time encoder; at least two transmit beamformers each receiving an output from a respective transmit processor and applying a transmit beamforming weight thereto; a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by the multiple transmit antennae.

[0020] Preferably, a delay of $\Delta\tau$ is interposed between one of the transmit processor outputs and a beamformer to delay the signal output from the transmit processor by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto, wherein the transmit processors do not add cyclic prefixes.

[0021] Conveniently, a delay of ψ is interposed between one of thetransmit processor outputs and a beamformer to delay the signal output from the transmit processor by ψ before the respective selected transmit beamforming weight is applied thereto, the transmit processors having a cyclic prefix length of $\Delta\psi$.

[0022] Advantageously, a processor to determine a power-delay-DOA profile estimate for channel h(n;k) is provided and, based on the profile, determine: the length, $\Delta\psi$ cyclic prefix to be added by the transmit processors; the delay ψ; and the transmit beamforming weights.

[0023] Conveniently, a processor is provided to estimate a downlink channel covariance matrix (DCCM) from the

uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

[0024] Preferably, the base station further comprises a first processor to determine a power-delay-DOA profile estimate for channel $h(n;k)$; and, based on the profile, determine: the length, $\Delta\psi$, of the cyclic prefix to be added by the transmit processors; the delay ψ; and the transmit beamforming weights; and a second processor to estimate a downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

[0025] Conveniently, the transmit and receive processors are selected from the group consisting of: OFDM, CDMA and TDMA processors.

[0026] Advantageously, the communications system comprises the base station and a mobile terminal having a single receive antenna, a receive processor to produce an output signal and a space-time decoder to decode the output signal.

[0027] A further aspect of the present invention provides a method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with a single receive antenna, the method comprising the steps of: providing a signal to be transmitted $s(n)$; space-time encoding a signal to be transmitted $s(n)$ to produce at least two separate signals $s_1(n), s_2(n)$, each on a respective output; delaying one of the space-time encoded output signals by $\Delta\tau$; applying respective selected transmit beamforming weights to the delayed and undelayed signals; feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal for transmission; feeding the summed signal to each of the multiple transmit antennae for transmission; transmitting the summed signals over the physical channel $h(k)$ with two time-delayed rays $h_1(k)$, $h_2(k)$; receiving the major components of the transmitted signals at a single receive antenna at substantially the same time; and space-time decoding the received signal.

[0028] Preferably, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

[0029] Conveniently, the delay $\Delta\tau$ is derived from downlink channel information.

[0030] A further aspect of the present invention provides a base station with multiple transmit antennae for communicating with a mobile terminal having a single receive antenna over physical channel $h(k)$ having two time-delayed rays $h_1(k)$, $h_2(k)$, the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal; at least two transmit beamformers each receiving an output from the space-time encoder and applying a transmit beamforming weight thereto; a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by each of the multiple transmit antennae, wherein a delay of $\Delta\tau$ is interposed between the space-time encoder and one of the beamformers such that the major components of the transmitted signals are received at a single receive antenna at substantially the same time.

[0031] Preferably, the communications system comprises the base station and a mobile terminal having a single receive antenna and a space-time decoder to decode the received signal.

[0032] One aim of the present invention is to seek to achieve, at the mobile terminal, diversity gain, beamforming gain as well as delay spread reduction simultaneously by using a base station with a multiple antenna array.

[0033] The advantages of the embodiments of the present invention are as follows:

- Beamforming gain and transmit diversity are achieved simultaneously;
- Based on power-delay-DOA profile, delay spread is reduced adaptively.
  - In two-ray environment, a frequency selective fading channel is transferred into a flat fading channel, yet the path diversity gain is maintained.
  - In hilly terrain (HT) environment, we can transfer a long delay spread channel into a short delay spread channel, yet still maintain the path diversity gain.
- With delay spread reduction and combined beamforming and transmit diversity, the invented systems provide high spectrum efficiency, yet consumes less transmission power.
- The invented systems also employ adaptive modulation to further improve the spectrum efficiency based on the diversity order and channel conditions.
- The mobile terminal is usually limited by physical size and battery power. The invented systems put the complicated processing at the base station, rather at the mobile terminal. Thus the mobile terminal complexity is reduced.
- The invented systems are well applicable for the applications which require high data rate for downlink transmission.

These applications include, for example, high speed downlink packet access (HSDPA) in 3rd generation partnership project (3GPP), wireless internet, and wireless multimedia communications.

[0034] In order that the present invention may be more readily understood, embodiments thereof will now be described, by way of example, with reference to the accompanying drawings, in which:

Figure 1 (*Prior Art*) is a schematic diagram illustrating Alamouti's permutation transmit diversity method;

Figure 2 is a schematic diagram illustrating a method embodying the present invention using transmit diversity with pre-equalization for frequency selective fading channels;

Figure 3 *(Prior Art)* is a schematic diagram illustrating orthogonal frequency division multiplexing (OFDM) with transmit diversity at: (a) a transmitter; and (b) a receiver;

Figure 4 *(Prior Art)* is a schematic diagram illustrating OFDM combined beamforming and transmit diversity for flat fading channels;

Figure 5 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming and transmit diversity at: (a) a transmitter; and (b) a receiver;

Figure 6 is a schematic diagram illustrating a method embodying the present invention using combined beamforming and transmit diversity for two ray (TR) frequency selective fading channels at (a) a transmitter; and (b) a receiver;

Figure 7 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming and transmit diversity for two ray (TR) models at: (a) a transmitter; and (b) a receiver;

Figure 8 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming and transmit diversity for hilly-terrain (HR) models at (a) a transmitter; and (b) a receiver; and

Figure 9 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming, transmit diversity and adaptive delay spread reduction: at (a) a transmitter; and (b) a receiver.

DETAILED DESCRIPTION OF THE INVENTION

[0035] The present invention revolves around the use of multiple antennas at the base station to improve the downlink performance of a wireless communication system. Downlink beamforming is effective in limiting interference pollution, which is of critical importance especially in multimedia communications. Transmit diversity is a powerful technique when receive diversity is impractical, especially for mobile terminals with size and/or power limitations. It can also be used to further improve downlink performance even though receive diversity is available.

[0036] In a multipath propagation environment, a receiver acquires several time-delayed, amplitude-scaled and direction of arrival (DOA) dependent versions of a transmitted signal. When the maximum time delay between the first-arrived and last-arrived versions of a signal along the various paths is smaller than the symbol interval, these paths are not resolvable in the time domain. However, these paths are resolvable in the spatial domain as they may come from different DOAs. Since each path may experience independent fading, using a beamforming antenna array, one obtains several independent channels, to which transmit diversity is applicable.

[0037] When the maximum relative delay is greater than the symbol interval, a frequency selective fading channel is observed. Frequency selectivity is beneficial for achieving diversity, however, it also yields inter-symbol interference (ISI) which needs to be suppressed at the receiver. This phenomenon becomes more and more prevalent as the data transmission rate increases. One way to suppress ISI is to use equalization at the receiver. The performance of an equalizer, however, depends on the frequency responses of the wireless channels. Specifically, when the channel's frequency responses have deep nulls in a certain frequency band, the equalization output yields noise enhancement, the effect of which can degrade the diversity gain obtained by the frequency selectivity. On the other hand, An adaptive equalizer often promotes error propagation problems when decision-directed symbols are used as reference signals, and the complexity of the equalizer is further complicated if the delay spread is large.

[0038] Another method of reducing ISI is to reduce the delay spread using adaptive antennas at the base station. For example, if the base station knows the direction-of-arrival (DOA) information of each delayed version of the received signal, it can then form a beam to one path whilst arranging for nulls or small antenna gains at the DOAs of the other paths. In this manner, the mobile terminal only receives one path of each transmitted signal. This method, though

simple in signal detection, sacrifices the diversity gain since use is only being made of one path.

[0039] Compared to receive diversity, transmit diversity has received greater attention during the past decade. Delay diversity as disclosed in A. Wittneben, "A new bandwidth efficient transmit antenna modulation diversity scheme for linear digital modulation", Proc. Of ICC'93, pp. 1630-1634, 1993, is one early transmit diversity technique using multiple transmit antennas. This method transforms a flat fading channel into a frequency selective fading channel making use of frequency diversity. An equalizer is provided at the mobile terminal in order to compensate for the artificially induced ISI. The performance of the equalizer depends on the frequency property of the channels. Further, an adaptive equalizer often promotes error propagation problems when decision-directed symbols are used as reference signals. In fact, it is shown in Y.C. Liang, Y. Li and K.J.R. Liu, "Feasibility of transmit diversity for IS-136 TDMA systems", Proc. Of VTC '98, pp. 2321-2324, 1998, that when the maximum Doppler frequency is over 40 Hz, this diversity method is even worse than that without diversity. In S.M. Alamouti, "A simple transmit diversity technique for wireless communications", IEEE Journal of Selected Areas in Communications, Vol.16, No.8, pp.1451-1458, October 1998, Alamouti proposed a permutation diversity method, whose performance is similar to maximal-ratio combining (MRC) receive diversity. This method only requires a simple receiver structure. More general transmit diversity methods are referred to as space-time coding methods as disclosed in V. Tarokh, N. Seshadri and A.R. Calderbank, "Space-time codes for high data rate wireless communication: Performance analysis and code construction", IEEE trans. On Information Theory, vol. 44, No. 3, pp. 744-765, March 1998. Space-time codes include space-time trellis codes (STTC) and space-time block codes (STBC). In fact, permutation diversity is the simplest class of STBC.

[0040] Figure 1 illustrating Alamouti's permutation diversity method shows the permutation diversity method with two transmit antennas 1, 2 equipped at the base station (BS). The signal $s(n)$ to be transmitted is first coded in a space-time coding module 3 The space-time coding module 3 works in the following way. It has one input port and two output ports. The input port accepts the transmitted sequence, s(0), s(1), ... The two output ports provide, in response, respective output signals $s_1(t)$ and $s_2(t)$ at time instants $t=n$ and $t=n+1$, where n is an even integer, as follows.

	$t=n$	$t=n+1$
$s_1(t)$	$s(n)/\sqrt{2}$	$s^*(n+1)/\sqrt{2}$
$s_2(t)$	$s(n+1)/\sqrt{2}$	$-s^*(n)/\sqrt{2}$

[0041] At a single receive antenna 4 at the mobile terminal the signals received at time instants $t=n$ and $t=n+1$ are given by

$$x(n) = \alpha_1 s_1(n) + \alpha_2 s_2(n) + w(n) \qquad (1)$$

$$x(n+1) = \alpha_1 s_1(n+1)+\alpha_2 s_2(n+1)+w(n+1) \qquad (2)$$

where α_1 and α_2 are the respective channel responses from the two transmit antennas 1, 2 to the receiver antenna 4, respectively; $w(n)$ is additive white Gaussian noise (AWGN).

[0042] The received signal is subsequently decoded by the space-time decoding module as follows. Specifically, equations (1) and (2) can be written in matrix forms:

$$\begin{bmatrix} x(n) \\ x(n+1) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} s(n) & s(n+1) \\ s^*(n+1) & -s^*(n) \end{bmatrix} \begin{bmatrix} \alpha_1 \\ \alpha_2 \end{bmatrix} + \begin{bmatrix} w(n) \\ w(n+1) \end{bmatrix} \qquad (3)$$

$$\begin{bmatrix} x(n) \\ x^*(n+1) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \alpha_1 & \alpha_2 \\ -\alpha_2^* & \alpha_1^* \end{bmatrix} \begin{bmatrix} s(n) \\ s(n+1) \end{bmatrix} + \begin{bmatrix} w(n) \\ w^*(n+1) \end{bmatrix} \qquad (4)$$

[0043] Therefore, channel coefficients can be estimated via equation (3) using training symbols; while equation (4) can be used for signal estimation/detection. This signal detection method is also called permutation decoding.

[0044] It is pointed out that, as opposed to delay diversity techniques which require a complicated equalizer at the receiver, the channel estimation and signal detection for permutation diversity involves very simple numerical operations. Also, compared to a one-transmitter/two-receiver receive diversity technique, even though the permutation diversity method has a 3 dB performance loss, it achieves the same order of diversity gain as receive diversity techniques using a maximal ratio combining (MRC) approach.

[0045] Permutation diversity can be extended to space-time block codes (STBC) and space-time trellis codes (STTC). All these codes achieve transmit diversity for flat fading environment.

[0046] One example of the invention applies Alamouti's diversity method to frequency selective fading channels. When the delay spread is greater than the symbol interval, frequency selective fading channels are observed. Figure 2 illustrates the system model applying Alamouti's diversity method to frequency selective fading channels. The transmitted signal, $s(n)$, is first coded using Alamouti's codes in the coding module 3, with the two branch outputs as $s_1(n)$ and $s_2(n)$. $s_1(n)$ and $s_2(n)$ are then passed into two pre-equalizers, 6, 7 having functions $g_1(k)$ and $g_2(k)$, to produce two output sequences $y_1(n)$ and $y_2(n)$. $y_1(n)$ and $y_2(n)$ are finally modulated and up-converted as RF signals, which are sent out through the transmit antennas 1, 2 as physical channels $h_1(k)$ and $h_2(k)$.

[0047] The functions $g_1(k)$ and $g_2(k)$ of the pre-equalizers 6,7 are used to pre-equalize the two physical channels, $h_1(k)$ and $h_2(k)$, respectively. By designing the pre-equalizers with zero-forcing criterion, the overall channel responses, $g_1(k) * h_1(k)$ and $g_2(k) * h_2(k)$, are now flat fading channels, with which Alamouti's coding/decoding method can be used. Here, " * " denotes a convolution operation.

[0048] In order to design the pre-equalizers 6,7, the real channel coefficients, $h_1(k)$ and $h_2(k)$, should be known at the base station/transmit antennas 1, 2. This can be done in two ways. For time-division duplex (TDD) systems, downlink channel coefficients are the same as uplink channel coefficients, which are derivable from the uplink using training symbols or blind techniques (up to a constant scaler). For frequency-division duplex (FDD) systems, the base station sends a set of training symbols to the mobile terminal, which then estimates and feeds back the downlink channel information to the base station.

[0049] The above methods are also applicable for other space-time codes.

[0050] Orthogonal frequency division multiplexing (OFDM) is a known and effective method of combatting the large delay spread problem. The combination of OFDM with a transmit diversity method not only suppresses large delay spread, but also achieves transmit diversity gain. Figure 3 shows a prior art OFDM system with two-antenna transmit diversity as described in Y. Li, N. Seshadri and S. Ariyavisitakul, "Channel estimation for OFDM systems with transmitter diversity in mobile wireless channels", IEEE Journal of Selected Areas in Communications, vol. 17, No. 3, pp. 461-471, March 1999. The signal to be transmitted, $S(n;k)$, is first coded using space-time codes in coding module 3, yielding two branch outputs as $S_1(n;k)$ and $S_2(n;k)$. $S_1(n;k)$ and $S_2(n;k)$ are then passed into respective normal OFDM transmit processors 8, 9, whose outputs are finally modulated and up-converted as RF signals, which are sent out through transmit antennas 1, 2.

[0051] At the single antenna receiver 4 at the mobile station, the received signal is passed into a normal OFDM receive processor 10, followed by a space-time decoder module 5. Specifically, the fast Fourier transform (FFT) output becomes

$$X(n;k) = H_1(n;k)S_1(n;k) + H_2(n;k)S_2(n;k) + W(n;k) \tag{5}$$

$$X(n;k+1) = H_1(n;k+1)S_1(n;k+1) + H_2(n;k+1)S_2(n;k+1) + W(n;k+1) \tag{6}$$

[0052] In (5) and (6), $H_1(n;k)$ and $H_2(n;k)$ are, respectively, the Fourier transforms of the channel impulse responses, $h_1(n;k)$ between transmit antenna 1 and receive antenna 4, and $h_2(n;k)$ between transmit antenna 2 and receive antenna 4; $W(n;k)$ is the FFT output of the additive noise, $w(n;k)$, received at the receive antenna 4.

[0053] Permutation decoding methods can be easily applied if $S_1(n;t)$ and $S_2(n;t)$ at time instants $t=k$ and $t=k+1$, where k is an even integer, are chosen as follows:

	$t=k$	$t=k+1$
$S_1(n;t)$	$S(n;k)\sqrt{2}$	$S^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$S(n;k+1)/\sqrt{2}$	$-S^*(n;k)/\sqrt{2}$

Prior Art: Combined beamforming and transmit diversity for flat fading channels.

[0054] The above three methods (Alamouti's permutation diversity method, a diversity method applied to frequency selective fading channels and OFDM with transmit diversity) achieve transmit diversity gain for flat fading channels, or frequency selective fading channels. The transmit antennas belong to diversity antennas, i.e., the antenna spacing is large, e.g., ten times wavelength, typically.

[0055] Figure 4 shows a known system combining beamforming and transmit diversity for flat fading channels as disclosed in R. Negi, A.M. Tehrani and J. Cioffi, "Adaptive antennas for space-time coding over block invariant multipath fading channels", Proc. of IEEE VTC, pp. 70-74, 1999. The signal to be transmitted, $s(n)$, is first coded using a space-time coder module 3, yielding two branch outputs as $s_1(n)$ and $s_2(n)$. $s_1(n)$ and $s_2(n)$ are then passed into two transmit beamformers 11,12, w_1 and w_2, respectively, followed by a signal combiner 13 which performs a simple summing function of the two inputs to producing a signal $x(n)$ for transmission which, in vector form, is as follows:

$$x(n) = w_1^H s_1(n) + w_2^H s_2(n) \tag{7}$$

[0056] To obtain spatial selectivity, the antenna spacing, d, is set to be small, e.g., half wavelength, and the number of transmit antennas 1A, 1B, 2, M, is greater than two. This is a beamforming antenna array, instead of a diversity antenna array. Suppose the physical channel consists of L spatially separated paths, whose fading coefficients and DOAs are denoted as $(\alpha_k(t),\theta_k)$, for $k = 1,\cdots,L$. If the maximum time delay relative to the first arrived path is smaller than the symbol interval, a flat fading channel is observed, and the instantaneous channel response, $h_d(t)$, can be expressed as follows:

$$h_d(t) = \sum_{k=1}^{L} \alpha_k(t) a_d(\theta_k) \tag{8}$$

where $a_d(\theta_k)$ is the downlink steering vector at DOA θ_k. The received signal, $y(n)$, at the mobile terminal is given by

$$y(n) = w_1^H h_d(t) s_1(n) + w_2^H h_d(t) s_2(n) + w(n) \tag{9}$$

[0057] By denoting $\beta_1(t)=w_1^H h_d(t)$, $\beta_2(t)=w_2^H h_d(t)$, the transmit beamforming weights can be estimated by maximizing the cost function:

$$J = E|\beta_1(t)|^2 + E|\beta_2(t)|^2 \tag{10}$$

$$\text{s.t.} \quad E[\beta_1(t)\beta_2^*(t)] = 0 \tag{11}$$

[0058] Maximum average signal to noise ratio (SNR) is obtained by maximising (10); while condition (11) guarantees that $\beta_1(t)$ and $\beta_2(t)$ are statistically uncorrelated, thus maximum diversity gain can be achieved.

[0059] Comparing (9) with (1), with the aid of downlink beamforming, two statistical uncorrelated fading channels, $\beta_1(t)$ and $\beta_2(t)$ have been artificially generated, with which space-time decoding can be used to recover the transmitted signal, $s(n)$. For Alamouti's diversity method, permutation decoding is applied.

[0060] The optimal transmit beamforming weight vectors are the eigenvectors corresponding to the two largest eigenvalues of the downlink channel covariance matrix (DCCM):

$$R_d = E\left[h_d(t)h_d^H(t)\right] \quad (12)$$

where the expectation is conducted over all fading coefficients. Suppose all paths have the same average power, or $E|\alpha_k(t)|^2 = 1/L$, the DCCM is given by

$$R_d = \frac{1}{L}\sum_{k=1}^{L} a_d(\theta_k)a_d^H(\theta_k) \quad (13)$$

[0061] For TDD, DCCM is the same as uplink channel covariance matrix (UCCM). For FDD, there are two ways to estimate the DCCM, both of which are based on the fact that uplink and downlink signals go through the same DOAs. The first method estimates the DOAs of all paths from the received uplink signals first, then constructs the downlink steering vectors, $a_d(\theta_k)$'s, and further DCCM R_d via equation (13). The second method estimates DCCM from UCCM directly via frequency calibration processing as disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4. This method does not involve DOA estimation and its associates and is therefore simple to implement.

[0062] This system achieves diversity gain and beamforming gain simultaneously for flat fading environment but it is desirable to extend that system into a frequency selective fading environment.

[0063] For mobile wireless communications without beamforming, the two ray (TR) model, typical urban (TU) model, and hilly terrain (HT) model are three commonly used power-delay profiles. When downlink beamforming is added, a power-delay-DOA profile should be considered. In picocell, microcell, and macrocell with TU model, there is less correlation between path delays and the DOAs. However, in macrocell with TR and HR models, the path delays are usually statistically dependent on the DOAs. We will show that for different environments, there exist different schemes to achieve combined beamforming and transmit diversity gains, as well as maximum spectrum efficiency.

[0064] Another example of the invention utilises OFDM to obtain combined beamforming and transmit diversity.

[0065] Combined beamforming and transmit diversity can be achieved by using OFDM for frequency selective fading channels. Figure 5 shows the OFDM system with combined beamforming and transmit diversity. Even though OFDM is selected as one example to show how the delay spread can be reduced, while yet maintaining beamforming and transmit diversity gain, other examples being other multi-carrier modulation schemes, such as MC-CDMA, MC-DS-CDMA and single carrier systems with cyclic prefix.

[0066] The transmitted signal at the *k*th tone of the *n*th block, $S(n;k)$, is first coded at the base station using space-time codes in coding module 3, yielding two branch outputs, $S_1(n;k)$ and $S_2(n;k)$. $S_1(n;k)$ and $S_2(n;k)$ are passed into respective normal OFDM transmit processors 8,9, followed by two transmit beamformers, 10,11, (w_1 and w_2) respectively. The beamforming outputs are finally combined in a combiner 13, and transmitted out through the transmit antennas 1A, 1B, 2 of the base station antenna array.

[0067] With the base station antenna array 1A, 1B, 2, the complex baseband representation of a wireless channel impulse response can be described as the following vector form

$$h_d(t;\tau) = \sum_m \sum_l \gamma_{m,l}(t)a_d(\theta_{m,l})\delta(\tau - \tau_m) \quad (14)$$

where τ_m is the delay of the *m*th path resolved in time, $\gamma_{m,l}(t)$ and $a_d(\theta_{m,l})$ are the complex amplitude and downlink steering vector corresponding to lth DOA of the *m*th delay path. Because of the motion of the vehicular, $\gamma_{m,l}(t)$'s are wide-sense stationary (WSS) narrow band complex Gaussian processes, which are zero-mean and statistically independent for different *m*'s, or *l*'s. Suppose all $\gamma_{m,l}(t)$'s have the same normalized correlation function, $r(t)$ ($r(0)=1$), but possibly different average power, σ_{ml}^2, then

$$E\left[\gamma_{m,l}(t+\Delta t)\gamma^{*}_{m,l}(t)\right]=\sigma^{2}_{m,l}r(\Delta t) \quad (15)$$

[0068] The Fourier transform (FT) of $h(t;\tau)$ at time instant t is given by

$$H_d(t;f)=\int_{\infty}^{\infty}h_d(t;\tau)e^{-j2\pi f\tau}d\tau=\sum_m\sum_l\gamma_{m,l}(t)a_d(\theta_{m,l})e^{-j2\pi f\tau_m} \quad (16)$$

[0069] For an OFDM system with block length T_b and tone spacing f_t, the discrete value of $H(t;f)$ is given by

$$H_d[n;k]\underline{\underline{\Delta}}H_d(nT_b;kf_t)=\sum_m\sum_l\gamma_{m,l}(nT_s)a_d(\theta_{m,l})e^{-j2\pi kf_t\tau_m} \quad (17)$$

thus the correlation function matrix of the frequency response for different times and frequencies is given by

$$r_d[\Delta n;\Delta k]=E\left[H_d[n+\Delta n;k+\Delta k]H_d^H[n;k]\right]=r(\Delta nT_b)\sum_m e^{-j2\pi\Delta kf_t\tau_m}R_{d,m} \quad (18)$$

where

$$R_{d,m}=\sum_l\sigma^2_{m,l}a_d(\theta_{m,l})a_d^H(\theta_{m,l})$$

is the downlink channel covariance matrix corresponding to the mth delay path. Note for Δn = 0 and Δk = 0,

$$r_d[0;0]=\sum_m\sum_l\sigma^2_{m,l}a_d(\theta_{m,l})a_d^H(\theta_{m,l})\underline{\underline{\Delta}}R_d \quad (19)$$

[0070] At the mobile terminal single antenna 4, the received signals are first passed into normal OFDM receive processor 10, followed by a permutation decoder 5. Within the normal OFDM receive processor, the FFT output becomes

$$X[n;k]=w_1^H\,H_d[n;k]S_1[n;k]+w_2^H H_d[n;k]S_2[n;k]+W[n;k] \quad (20)$$

$$X[n;k+1] = w_1^H H_d[n;k+1]S_1[n;k+1]+w_2^H H_d[n;k+1]S_2[n;k+1]+W[n;k+1] \quad (21)$$

where $W[n;K]$ is zero mean AWGN.

[0071] By denoting $\beta_1 = w_1^H H_d[n;k]$, $\beta_2 = w_2^H H_d[n;k]$, the beamforming weights can be estimated by maximizing the cost function:

$$J = E|\beta_1|^2 + E|\beta_2|^2 \quad (22)$$

$$\text{s.t.} \quad E[\beta_1\beta_2^*] = 0 \quad (23)$$

[0072] Again, maximum average SNR is obtained through maximizing equation (22); while condition (23) guarantees that β_1 and β_2 are statistically uncorrelated, thus maximum diversity gain can be achieved.

[0073] The optimal transmit beamforming weight vectors are the eigenvectors corresponding to the two largest eigenvalues of downlink channel covariance matrix (DCCM) R_d.

$$R_d = E[H_d[n;k]H_d^H[n;k]] \quad (24)$$

[0074] Comparing equations (20) and (21) with equations (5) and (6), with the aid of downlink beamforming, two uncorrelated fading channels are generated, with which the space-time decoding can be used to recover the transmitted signal. Permutation decoding method can be applied if $S_1(n;k)$ and $S_2(n;k)$ are chosen as follows.

	$t=k$	$t=k+1$
$S_1(n;t)$	$s(n;k)/\sqrt{2}$	$s^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$s(n;k+1)/\sqrt{2}$	$-s^*(n;k)/\sqrt{2}$

[0075] A frequency calibration method for DCCM estimation for OFDM.

[0076] In order to generate the downlink beamforming weights, it is first necessary to construct the DCCM. A frequency calibration (FC) method disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is applied.

[0077] Using a similar method, we can show that the correlation function matrix of the uplink frequency response for different times and frequencies is given by

$$r_u[\Delta n;\Delta k] = E[H_u[n+\Delta n;k+\Delta k]H_u^H[n;k]] = r(\Delta n T_b)\sum_m e^{-j2\pi\Delta k f_t \tau_m} R_{u,m} \quad (25)$$

where

$$R_{u,m} = \sum_l \sigma_{m,l}^2 a_u(\theta_{m,l}) a_u^H(\theta_{m,l})$$

is the uplink channel covariance matrix corresponding to the mth delay path. Note for Δn = 0 and Δk = 0,

$$r_u[0;0] = \sum_m \sum_l \sigma_{m,l}^2 a_u(\theta_{m,l}) a_u^H(\theta_{m,l}) \underline{\underline{\Delta}} R_u \tag{26}$$

[0078] Comparing equations (19) and (26), the FC method devised in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is used to estimate the DCCM from UCCM.

[0079] This system provides diversity gain and beamforming gain for OFDM systems. In this system, the length of cyclic prefix is determined by the maximum physical time delay, and is the same as that in a normal OFDM system. Thus it is readily applicable to the environment in which the DOA is statistically independent of the time delay.

[0080] When the DOA of a path is statistically related to the path delay, e.g., in TR and HR environments, one can not only achieve beamforming gain and diversity gain simultaneously, but also reduce the cyclic prefix, thus obtaining improved spectrum efficiency.

[0081] A further example of the present invention utilises combined beamforming and transmit diversity for frequency selective fading channels for two ray (TR) models.

[0082] Suppose the physical channel follows a TR model. With the base station antenna array, the complex baseband representation of a wireless channel impulse response can be described as the following vector form

$$h_d(t;\tau) = \sum_{m=1}^{2} h_{d,m}(t)\delta(\tau - \tau_m) \tag{27}$$

with

$$h_{d,m}(t) = \sum_l \gamma_{m,l}(t) a_d(\theta_{m,l}) \tag{28}$$

where τ_m is the delay of the mth path resolved in time, $\gamma_{m,l}(t)$ and $a_d(\theta_{m,l})$ are the complex amplitude and downlink steering vector corresponding to lth DOA of the mth delay path. Because of the motion of the vehicular, $\gamma_{m,l}(t)$'s are wide-sense stationary (WSS) narrow band complex Gaussian processes, which are zero-mean and statistically independent for different m's, or l's. Suppose all $\gamma_{m,l}(t)$'s have the same normalized correlation function, $r(t)$ ($r(0) = 1$), but possibly different average power, $\sigma_{m,l}^2$, then

$$E\left[\gamma_{m,l}(t+\Delta t)\gamma_{m,l}^*(t)\right] = \sigma_{m,l}^2 r(\Delta t) \tag{29}$$

[0083] ISI exists when $\Delta\tau = \tau_2 - \tau_1$ is greater than the symbol interval. With combined beamforming and diversity technique, if the two rays are spatially separated, it is possible to transfer a frequency selective fading channel into a flat fading channel, yet maintain the transmit diversity.

[0084] Figure 6 shows a communication system with combined beamforming and transmit diversity for two-ray frequency selective fading channels. The signal to be transmitted, $s(n)$, is first coded in a coding module 3 using space-time codes, with the two branch outputs as $s_1(n)$ and $s_2(n)$. $s_1(n)$ is then fed through a delay 14 to delay s(n) by $\Delta\tau$, yielding $x_1(n)$, which is further passed to transmit beamformer 11, (w_1). The second branch output $s_2(n)$ is directly passed to the other transmit beamformer 12, (w_2). The beamforming outputs are then combined in combiner 13 and sent by transmit antennas 1A, 1B, 2, yielding the transmitted signal as follows:

$$x(n) = w_1^H x_1(n) + w_2^H s_2(n) \tag{30}$$

[0085] The received signal, *y(n),* at the mobile terminal single antenna 4 is given by

$$\begin{aligned} y(n) &= w_1^H h_{d,1} x_1(n) + w_1^H h_{d,2} x_1(n - \Delta\tau) \\ &+ w_2^H h_{d,1} s_2(n) + w_2^H h_{d,2} s_2(n-\Delta\tau) + w(n) \end{aligned} \tag{31}$$

[0086] Denoting $z(n) = y(n+\Delta\tau)$, and considering the pre-alignment of the two transmitted signals, gives:

$$\begin{aligned} z(n) &= w_1^H h_{d,1} s_1(n) + w_1^H h_{d,2} s_1(n-\Delta\tau) \\ &+ w_2^H h_{d,1} s_2(n + \Delta\tau) + w_2^H h_{d,2} s_2(n) + w(n + \Delta\tau) \end{aligned} \tag{32}$$

[0087] The beamforming weights are chosen such that the first branch output, $s_1(n)$, just goes through the first path, $h_{d,1}$ between the base station antenna array and the receive antenna 4; while the second branch output, $s_2(n)$, just goes through the second path, $h_{d,2}$ between the base station antenna array and the receive antenna 4. Mathematically,

$$\begin{cases} w_1^H h_{d,2} = 0 \\ \left| w_1^H h_{d,1} \right|^2 = \max \end{cases}$$

and

$$\begin{cases} w_2^H h_{d,1} = 0 \\ \left| w_2^H h_{d,2} \right|^2 = \max \end{cases}$$

[0088] In this case the ISI terms are suppressed completely, and *z(n)* can be written as

$$z(n) = w_1^H h_{d,1} s_1(n) + w_2^H h_{d,2} s_2(n) + w(n + \Delta\tau) \tag{33}$$

[0089] Thus the frequency selective fading channel is now transformed into a flat fading channel, with which the transmit diversity method can be applied.

[0090] Conveniently, the transmit beamforming weights can be chosen by maximizing the average transmit SINR functions:

$$J_1(w_1) = \frac{w_1^H R_{d,1} w_1}{w_1^H R_{d,2} w_1} \text{ and } J_2(w_1) = \frac{w_2^H R_{d,2} w_2}{w_2^H R_{d,1} w_2}$$

where

$$R_{d,m} = E\left[h_{d,m}(t)h_{d,m}^{H}(t)\right] = \sum_{l}\sigma_{m,l}^{2}a_{d}(\theta_{m,l})a_{d}^{H}(\theta_{m,l}) \tag{34}$$

is the downlink channel covariance matrix of the mth path.

[0091] Preferably, the transmit beamforming weights can be chosen by maximizing the average receive SINR at the mobile receiver, i.e.,

$$J = \frac{w_1^H R_{d,1} w_1 + w_2^H R_{d,2} w_2}{w_1^H R_{d,2} w_1 + w_2^H R_{d,1} w_2 + \sigma_n^2}, \tag{35}$$

[0092] Advantageously, the transmit beamforming weights, w_m, can be chosen as the principal eigenvector of $R_{d,m}$.

[0093] Again, the frequency calibration method disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is used to estimate the DCCM from UCCM directly.

[0094] The above method for achieving combined beamforming and transmit diversity gain is called pre-alignment (PAL) method. The purpose of delaying $s_1(n)$ by $\Delta\tau$ is to make sure that the major components of the two sequences, $s_1(n)$ and $s_2(n)$ arrive at the receiver at the same time. Therefore, the delay spread has been reduced to zero. On the other hand, beamforming is used to minimize the ISI effect as well as to artificially generate two uncorrelated channels, with which the transmit diversity gain is achieved.

[0095] The PAL method requires the delay information, $\Delta\tau$, which is embedded in the downlink power-delay-DOA (PDD) profile. Even though the PDD profile is time varying, it changes slowly in time. Also, downlink PDD profile is almost the same as uplink PDD profile, which can be estimated from received uplink signals.

[0096] The PAL method can also be applied to the systems whose number of rays is grater than 2. In this case, it requires more than 2 branches of space-time coding outputs, and each output except the first one corresponds to one delay.

[0097] If the number of space-time coding outputs is fixed, say 2, the two major rays can be selected in order to generate the delay, $\Delta\tau$, and the transmit beamforming weights. The direct application of this system is to reduce inter-fmger-interference in CDMA as the total number of fingers is reduced.

[0098] Conventionally, when the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

[0099] Advantageously, when the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average transmit SINR function at the base station is maximized for each ray.

[0100] Preferably, when the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average receive SINR function at the mobile terminal is maximized.

[0101] Another example of the present invention utlises OFDM with combined beamforming and transmit diversity for frequency selective fading channels for two ray (TR) models.

[0102] There is a direct use of delay spread reduction in OFDM. In a typical OFDM system, a cyclic prefix is added in order to remove the ISI and to guarantee the orthogonality between each sub-channel. The length of the cyclic prefix should be greater than the maximum time delay, which can be as large as 40 µs for a mobile wireless communication environment. The adding of the cyclic prefix not only degrades the spectrum efficiency, but also occupies one portion of the transmit power. The spectrum efficiency and power efficiency of the OFDM system can be greatly improved if the cyclic prefix can be reduced while maintaining the same performance.

[0103] Suppose the physical channel follows a TR model with parameters $(\alpha_k,\theta_k,\tau_k)$, k = 1,2 and $\tau_1<\tau_2$. α_k's are statistically independent, zero mean complex Gaussian processes with variance σ_k^2. ISI exists when $\Delta\tau = \tau_2-\tau_1$ is greater than the inverse of bandwidth.

[0104] Figure 7 illustrates an OFDM system with combined beamforming and transmit diversity for TR models embodying the present invention. The transmitted signal at the kth tone of the nth block, $S(n;k)$, is first coded using space-time codes in coding module 3, yielding two branch outputs, $S_1(n;k)$ and $S_2(n;k)$. Both branch outputs $S_1(n;k)$ and $S_2(n;k)$ are passed into respective OFDM transmit processors 8,9 without adding cyclic prefixes. $S_1(n;k)$ is then delayed

in delay 14 by $\Delta\tau$, yielding $X_1(n;k)$, which is further passed to transmit beamformer 11, (w_1). The second branch output $S_2(n;k)$ is directly passed to the other transmit beamformer 12, (w_2). The beamforming outputs are then combined and sent on the base station transmit antenna array 1A, 1B, 2, yielding the transmitted signal as follows:

$$x(n;k) = w_1^H x_1(n;k) + w_2^H s_2(n; k) \qquad (36)$$

[0105] At the mobile terminal single antenna 4, the received signals are first passed into a normal OFDM receive processor 10. The beamforming weights are chosen such that the first branch output, $S_1(n;k)$ or its inverse FFT (IFFT), $s_1(n;k)$, just goes through the first path, $h_1(n;k)$ between the base station antenna array and the receive antenna 4; while the second branch output, $S_2(n;k)$ or its inverse FFT (IFFT), $s_2(n;k)$, just goes through the second path, $h_1(n;k)$ between the base station antenna array and the receive antenna 4. Once the transmit beamforming weights are properly chosen, the FFT output of the received signal at the mobile station becomes

$$Z[n;k] = w_1^H H_1[n;k]S_1[n;k] + w_2^H H_2[n;k]S_2[n;k] + W[n;k + \lfloor \Delta\tau f_t \rfloor]$$

(37)

[0106] Comparing equation (37) with equation (5), with the aid of downlink beamforming, two different channels have been artificially created which can be space-time decoded by module 5 to recover the transmitted signal. Further, permutation decoding method can be easily applied if $S_1(n;k)$ and $S_2(n;k)$ are chosen as follows.

	$t=k$	$t=k+1$
$S_1(n;t)$	$s(n;k) / \sqrt{2}$	$s^*(n;k+1) / \sqrt{2}$
$S_2(n;t)$	$s(n;k+1) / \sqrt{2}$	$-s^*(n;k) / \sqrt{2}$

[0107] When PAL is applied to an OFDM system with combined beamforming and transmit diversity for TR models, it is not necessary to add the cyclic prefix. Thus benefiting from the advantages of: transmit diversity; beamforming gain; and increased spectrum efficiency.

[0108] Conveniently, the transmit beamforming weights can be chosen by maximizing the average transmit SINR functions.

[0109] Preferably, the transmit beamforming weights can be chosen by maximizing the average receive SINR at the mobile receiver.

[0110] Advantageously, the transmit beamforming weights, w_m, can be chosen as the principal eigenvector of $R_{d,m}$.

[0111] Again, the frequency calibration method disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is used to estimate the DCCM from UCCM directly.

[0112] A comparison of the spectrum efficiency and power savings by using this delay spread reduction method will follow.

[0113] A further example of the invention utilises OFDM with combined beamforming and transmit diversity for frequency selective fading channels for hilly terrain (HT) models.

[0114] Even though the maximum time delay can be as large as 40 µs, a wireless channel satisfying HT model can be described by several dominated clustered paths, each of which has a small delay spread. These clustered paths are also spatially separated. For an OFDM with typical HT power-delay profile whose maximum time delay is 20 µs, and maximum delay spread for each clustered path is 2 µs, the minimum length of cyclic prefix is 20µs in order to remove the ISI. However, with the PAL method, the cyclic prefix duration can be reduced to 2 µs .

[0115] Suppose the two clustered paths are delayed by ψ,and for simplicity, assume the delay spread for each clustered path is $\Delta\psi$. The impulse response of the time varying channel can be described as

$$h(t;\tau)=h_1(t;\tau)[u(\tau)-u(\tau-\Delta\psi)]+h_2(t;\tau-\psi)[u(\tau-\psi)-u(\tau-\psi-\Delta\psi)] \tag{38}$$

where $h_1(t;\tau)$ and $h_2(t;\tau)$ correspond to the channel responses of the first and second clustered paths, respectively; and $u(x)$ is a unit step function..

[0116] Figure 8 shows an OFDM system embodying the present invention with combined beamforming and transmit diversity for hilly terrain (HT) model in encoder module 3. The signal to be transmitted at the kth tone of the nth block, $S(n;k)$, is first coded using space-time codes in encoder module 3, yielding two branch outputs, $S_1(n;k)$ and $S_2(n;k)$ which are passed into respective normal OFDM transmit processors 8,9, whose cyclic prefix length is $\Delta\psi$, instead of $\psi + \Delta\psi$. The output from the first branch is then delayed by ψ in delay 15, while the output from the second branch remains unchanged. After that, the signals are passed into respective transmit beamformers 11,12, (w_1 and w_2), respectively. The beamforming outputs are then combined in combiner 13, and transmitted out through the base station transmit antenna array 1A, 1B, 2.

[0117] The beamforming weights are chosen such that the first branch input just goes through the first clustered path, while the second branch input just goes through the second clustered path - i.e. the beamforming weights are chosen such that the first branch output, $s_1(n)$, just goes through the first path, $h_{d,1}$between the base station antenna array and the receive antenna 4; while the second branch output, $s_2(n)$, just goes through the second path, $h_{d,2}$between the base station antenna array and the receive antenna 4. The signals received at the mobile terminal single antenna 4 are first passed into a normal OFDM receive processor 10, followed by a space-time decoding module 5. Within the normal OFDM receive processor 10, the received signal after FFT becomes

$$Z[n;k]=w_1^H H_1[n;k]S_1[n;k]+w_2^H H_2[n;k]S_2[n;k]+W[n;k+\lfloor\psi f_t\rfloor] \tag{39}$$

where $\lfloor x\rfloor$ denotes the maximum integer which is not greater than x. Comparing equation (39) with equation (5), with the aid of downlink beamforming, two different channels have been artificially generated, which are space-time decoded to recover the transmitted signal. Permutation decoding methods can be easily applied if $S_1(n;k)$ and $S_2(n;k)$ are chosen as follows.

	$t=k$	$t=k+1$
$S_1(n;t)$	$s(n;k)/\sqrt{2}$	$s^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$s(n;k+1)/\sqrt{2}$	$-s^*(n;k)/\sqrt{2}$

[0118] Conveniently, the transmit beamforming weights can be chosen by maximizing the average transmit SINR functions.

[0119] Preferably, the transmit beamforming weights can be chosen by maximizing the average receive SINR at the mobile receiver.

[0120] Advantageously, the transmit beamforming weights, w_m, can be chosen as the principal eigenvector of $R_{d,m}$.

[0121] As previously mentioned, there follows a comparison the spectrum efficiency of a OFDM system with different cyclic prefix lengths.

[0122] The parameters are Bandwidth B = 800 kHz, maximum time delay = 40. For HT models, the maximum delay spread for each clustered path is 5. To make the tones orthogonal to each other, the symbol duration is N/B, where N is the number of tones in each OFDM symbol. The total block length is the summation of the symbol duration and the additional guard interval, which is 40, 5, and 0 for OFDM without PAL, HT with PAL and TR with PAL, respectively.

[0123] Table I illustrates the uncoded transmit data rate for OFDM systems with different number of tones using QPSK modulation. It is seen that, for a given modulation scheme and with the same number of tones, the transmit data rate can increase to 1.6Mbps for TR environments by using PAL, independent of the N value. For HT with PAL, the spectrum efficiency is also increased as compared with that without PAL.

Table I:

transmit data rate comparison			
	N=128	N=64	N=32
Without PAL	1.28 Mbps	1.07 Mbps	800 kbps
HT with PAL	1.55 Mbps	1.51 Mbps	1.42 Mbps
TR with PAL	1.6 Mbps	1.6 Mbps	1.6 Mbps

[0124] Here follows a comparison of the power savings for OFDM with different lengths of cyclic prefix:

[0125] Due to the adding of a cyclic prefix, the effective $\frac{E_b}{N_0}$ is smaller than the actual transmit $\frac{E_b}{N_0}$. With delay spread reduction, the transmit power is more efficiently used. Table II illustrates the power savings for OFDM systems with delay spread reduction using PAL for different number of tones in each OFDM block, as compared to normal OFDM systems.

Table II:

Power savings			
	N=128	N=64	N=32
HT with PAL	0.84 dB	1.5 dB	2.5 dB
TR with PAL	0.97 dB	1.76 dB	3.0 dB

Beamforming and diversity gain:

[0126] With combined beamforming and diversity gain, it takes less $\frac{E_b}{N_0}$ in order for the system to achieve a given bit-error-rate (BER) requirement. Alternatively, the beamforming and diversity gain can be translated to larger spectrum efficiency using higher modulation scheme such as 128 QAM or 256 QAM.

[0127] A further embodiment of the present invention relates to adaptive delay spread reduction with combined beamforming and diversity gain:

[0128] The previously described embodiments are designed for different environments. In real applications, the power-delay-DOA (PDD) profile may change with respect to time due to the motion of a vehicle, thus the delay spread reduction scheme should follow this variation accordingly in order to achieve maximum spectrum efficiency. Figure 9 shows an OFDM system with combined beamforming, transmit diversity and adaptive delay spread reduction for downlink embodying the present invention. The OFDM system of Figure 9 comprises the system of Figure 8 but supplemented by UCCM estimation and power-delay-DOA profile estimation. Thus, in addition to the functionality provided by the system of Figure 8, this system has the following functionality.

- From uplink signals received at the base station, the time-delay and direction-of-arrival (DOA) information is estimated for each received path, using training sequences or blind techniques. Based on the estimated time-delay and DOA information, uplink power-delay-DOA (PDD) profile, and each clustered path's UCCM are estimated;
- Based on uplink PDD profile, the following parameters are determined: diversity order, time delays for each clustered path, and the maximum delay spread for the clustered paths.
- The uplink PDD profile is used to design the adaptive delay reduction scheme, thus the adaptive cyclic prefix adding scheme;
- Each clustered path's DCCM is estimated from its corresponding UCCM using FC method disclosed in Y-C. Liang and F. Chin "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4, then applied, together with time delay information, for constructing transmit beamforming weights;
- The base station informs the MS the length of added cyclic prefix;
- Adaptive modulation is also used to further improve the spectrum efficiency based on the diversity order/channel condition. Specifically, based on uplink PDD profile, the maximum achievable diversity order is determined. If the achievable diversity order is large, a higher modulation scheme is applied; otherwise, a smaller modulation scheme is applied.

[0129] It should be noted that the number of branch outputs after space-time coding in module 3 can be greater than two, depending on the diversity order to be achieved.

[0130] The above description considers the combined beamforming, transmit diversity and delay spread reduction implemented at the base station. In fact, multiple diversity antennas can be added at the mobile terminal as well to achieve receive diversity. In this case, larger diversity gains can be achieved:

[0131] Even though OFDM is used to show how the delay spread can be reduced, while yet maintaining beamforming and transmit diversity gain, the disclosure in this application can be applied to other multi-carrier modulation schemes, such as MC-CDMA, MC-DS-CDMA and single carrier systems with cyclic prefix.

[0132] In a multiuser environment, the beamforming weights can be generated by considering all users' channel/ DOA information; therefore, the disclosure in this application is applicable in different multiple access schemes, such as time-division-multiple-access (TDMA), frequency-division-multiple-access (FDMA), and code-division-multiple-access (CDMA).

"comprising" means "including or consisting of".

[0133] The features disclosed in the foregoing description, or the following claims, or the accompanying drawings, expressed in their specific forms or in terms of a means for performing the disclosed function, or a method or process for attaining the disclosed result, as appropriate, may, separately, or in any combination of such features, be utilised for realising the invention in diverse forms thereof.

Claims

1. A method of achieving transmit diversity gain for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of:

   providing a signal to be transmitted $s(n)$;
   space-time encoding the signal $s(n)$ to produce at least two separate signals $s_1(n),s_2(n)$, each on a respective output;
   feeding each output signal $s_1(n),s_2(n)$ to a zero-forcing pre-equaliser having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$;
   feeding the output signal $x_1(n)$, $x_2(n)$ of each pre-equaliser to a transmit antenna;
   transmitting the output signals $x_1(n)$, $x_2(n)$ over respective physical channels $h_1(k)$, $h_2(k)$;
   receiving the output signals $x_1(n)$, $x_2(n)$ at at least a single receive antenna; and space-time decoding the received signals, wherein
   the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(s)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

2. A method according to Claim 1, wherein the communications system is a time-division duplex system and the method includes the further step of deriving the real channel coefficients from uplink channel coefficients for use in selecting the functions $g_1(k)$, $g_2(k)$ of the pre-equalisers.

3. A method according to Claim 2, wherein the step of deriving the real channel coefficients from uplink channel coefficients uses training symbols from the uplink channel.

4. A method according to Claim 2, wherein the step of deriving the real channel coefficients from uplink channel coefficients uses blind techniques.

5. A method according to Claim 1, wherein the communications system is a frequency-division duplex system and the method includes the further step of deriving the real channel coefficients by sending a set of training symbols to the receive antenna of the mobile terminal, the mobile terminal estimating the real channel coefficients and feeding back channel coefficient information to the base station.

6. A base station with multiple transmit antennae for communicating with a mobile terminal having at least a single receive antenna over physical channels $h_1(k)$, $h_2(k)$ the base station comprising:

   a space-time encoder having an input of a signal to be transmitted $s(n)$ and at least two outputs each producing a separate signal $s_1(n),s_2(n)$;
   at least two zero-forcing pre-equalisers, each fed by a respective output signal $s_1(n),s_2(n)$ and having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$; and
   at least two transmit antennae, each being fed by the output signal $x_1(n)$, $x_2(n)$ of a respective one of the pre-

equalisers, wherein the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)*h_1(k)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

7. A communications system comprising the base station of Claim 6 and a mobile terminal having at least a single receive antenna and a space-time decoder to decode the signals received from the base station.

8. A method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of:

providing a signal to be transmitted $S(n;k)$;
space-time encoding the signal $S(n;k)$ to produce at least two separate signals $S_1(n;k)$,$S_2(n;k)$, each on a respective output;
feeding each output signal $S_1(n;k)$,$S_2(n;k)$ to a transmit processor to produce an output signal $X_1(n;k)$, $X_2(n;k)$;
applying respective selected transmit beamforming weights to each output signal $X_1(n;k)$, $X_2(n;k)$;
feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal $X(n;k)$ for transmission;
feeding the summed signal $X(n;k)$ to each of the multiple transmit antennae for transmission;
transmitting the signals $X(n;k)$ over physical channel $h(n;k)$;
receiving the received signal $Y(n;k)$ at at least a single receive antenna;
feeding the received signal $Y(n;k)$ to a receive processor to produce an output signal; and
space-time decoding the received signal.

9. A method according to Claim 8, wherein the respective transmit beamforming weights are selected as the eigenvectors corresponding to the two largest eigenvalues of the downlink channel covariance matrix (DCCM) of the physical channels $h(n;k)$.

10. A method according to Claim 8, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the transmit processors do not add cyclic prefixes and one of the output signals from the transmit processors is delayed by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto.

11. A method according to Claim 8, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights being chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

12. A method according to Claim 8, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights being chosen such that the average transmit SINR function at the base station is maximized for each ray.

13. A method according to Claim 8, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights being chosen such that the average receive SINR function at the mobile terminal is maximized.

14. A method according to Claim 8, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights for each ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that ray.

15. A method according to Claim 8, wherein the physical channel $h(n;k)$ consists of two time-delayed clustered rays, $h_1(n;k)$ and $h_2(n;k)$, with delay ψ, and maximum excess delay for the clusters $\Delta\psi$, the transmit processors have a cyclic prefix length of $\Delta\psi$ and one of the output signals from the transmit processors is delayed by ψ before the respective selected transmit beamforming weight is applied thereto.

16. A method according to Claim 15, wherein the beamforming weights are chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple

transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

17. A method according to Claim 15, wherein the beamforming weights being chosen such that the average transmit SINR function at the base station is maximized for each clustered ray.

18. A method according to Claim 15, wherein the beamforming weights being chosen such that the average receive SINR function at the mobile terminal is maximized.

19. A method according to Claim 15, wherein the beamforming weights for each clustered ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that clustered ray.

20. A method according to Claim 15, comprising the further steps of:

estimating a power-delay-DOA profile for channel $h(n;k)$; and, based on the profile: determining the cyclic prefix, $\Delta\psi$, to be added by the transmit processors; determining the delay ψ; diversity order and modulation scheme; and determining the transmit beamforming weights.

21. A method according to Claim 20, comprising the further step of estimating the downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

22. A method according to Claim 21, comprising the further step of determining the diversity order and modulation scheme based on the profile.

23. A method according to Claim 8, wherein the transmit and receive processors are selected from the group consisting of: OFDM, MC-CDMA MC-DS-CDMA and a single carrier system with cyclic prefix.

24. A base station with multiple transmit antennae for communicating with a mobile terminal having at least a single receive antenna over physical channel $h(k)$, the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal;
at least two transmit processors each receiving one of the outputs from a respective space-time encoder;
at least two transmit beamformers each receiving an output from a respective transmit processor and applying a transmit beamforming weight thereto;
a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by the multiple transmit antennae.

25. A base station according to Claim 24, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, further comprising a delay of $\Delta\tau$ interposed between one of the multiple access transmit processor outputs and a beamformer to delay the signal output from the transmit processor by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto, wherein the transmit processors do not add cyclic prefixes.

26. A base station according to Claim 24, wherein the physical channel $h(n;k)$ consists of two time-delayed clustered rays, $h_1(n;k)$ and $h_2(n;k)$, with delay ψ and maximum excess delay for the clusters $\Delta\psi$, further comprising a delay of ψ interposed between one of the multiple access transmit processor outputs and a beamformer to delay the signal output from the transmit processor by ψ before the respective selected transmit beamforming weight is applied thereto, the transmit processors having a cyclic prefix length of $\Delta\psi$.

27. A base station according to Claim 24, further comprising a first processor to determine a power-delay-DOA profile estimate for channel $h(n;k)$; and, based on the profile, determine: the length, $\Delta\psi$, of the cyclic prefix to be added by the transmit processors; the delay ψ; diversity order and modulation scheme; and the transmit beamforming weights.

28. A base station according to Claim 27, further comprising a second processor to estimate a downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming

weights.

29. A base station according to Claim 15, wherein the transmit and receive processors are selected from the group consisting of: OFDM, MC-CDMA MC-DS-CDMA and single carrier system with cyclic prefix.

30. A communications system comprising the base station of Claim 24 and a mobile terminal having at least a single receive antenna, a receive processor to produce an output signal and a space-time decoder to decode the output signal.

31. A method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of:

providing a signal to be transmitted $s(n)$;
space-time encoding a signal to be transmitted $s(n)$ to produce at least two separate signals $s_1(n), s_2(n)$, each on a respective output;
delaying one of the space-time encoded output signals by $\Delta\tau$;
applying respective selected transmit beamforming weights to the delayed and undelayed signals;
feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal for transmission;
feeding the summed signal to each of the multiple transmit antennae for transmission;
transmitting the summed signals over the physical channel $h(k)$;
receiving the major components of the transmitted signals at at least a single receive antenna at substantially the same time; and
space-time decoding the received signal.

32. A method according to Claim 31, wherein the physical channel $h(k)$ consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

33. A method according to Claim 31, wherein the physical channel $h(k)$ consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights are chosen such that the average transmit SINR function at the base station is maximized for each ray.

34. A method according to Claim 31, wherein the physical channel $h(k)$ consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights are chosen such that the average receive SINR function at the mobile terminal is maximized.

35. A method according to Claim 31, wherein the physical channel $h(k)$ consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights for each ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that ray.

36. A method according to Claim 31, wherein the physical channel $h(k)$ consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the delay $\Delta\tau$ is derived from downlink channel information.

37. A method according to Claim 31, wherein the physical channel $h(k)$ consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the delay $\Delta\tau$ is derived from uplink channel information.

38. A method according to Claim 31, wherein the physical channel $h(k)$ consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

39. A method according to Claim 31, wherein the physical channel $h(k)$ consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average transmit SINR function at the base station is maximized for each ray.

40. A method according to Claim 31, wherein the physical channel *h*(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average receive SINR function at the mobile terminal is maximized.

41. A method according to Claim 31, wherein the physical channel *h*(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights for each ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that ray.

42. A method according to Claim 31, wherein the physical channel *h*(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the delay $\Delta\tau$ is derived from downlink channel information.

43. A method according to Claim 31, wherein the physical channel *h*(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the delay $\Delta\tau$ is derived from uplink channel information.

44. A base station with multiple transmit antennae for communicating with a mobile terminal having at least a single receive antenna over physical channel $h(k)$ having two time-delayed rays, $h_1(k)$ and $h_2(k)$, the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal;
at least two transmit beamformers each receiving an output from the space-time encoder and applying a transmit beamforming weight thereto;
a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by each of the multiple transmit antennae, wherein a delay of $\Delta\tau$ is interposed between the space-time encoder and one of the beamformers such that the major components of the transmitted signals are received at at least a single receive antenna at substantially the same time.

45. A communications system comprising the base station of Claim 24 and a mobile terminal having at least a single receive antenna and a space-time decoder to decode the received signal.

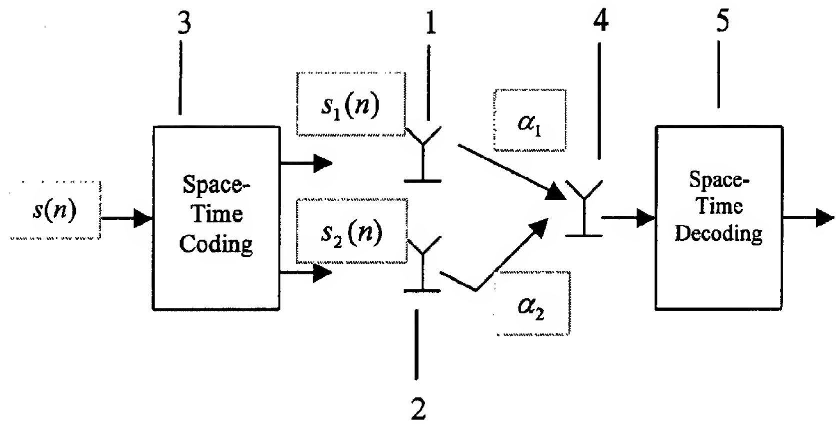

Figure1: Prior Art

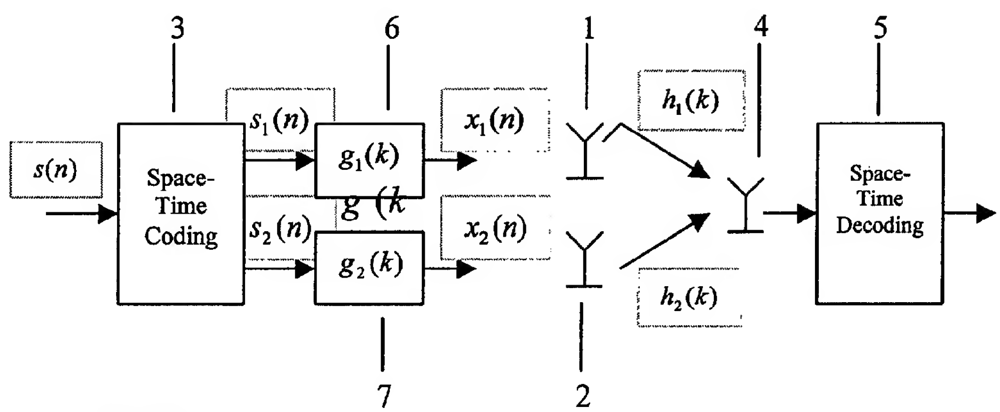

Figure2

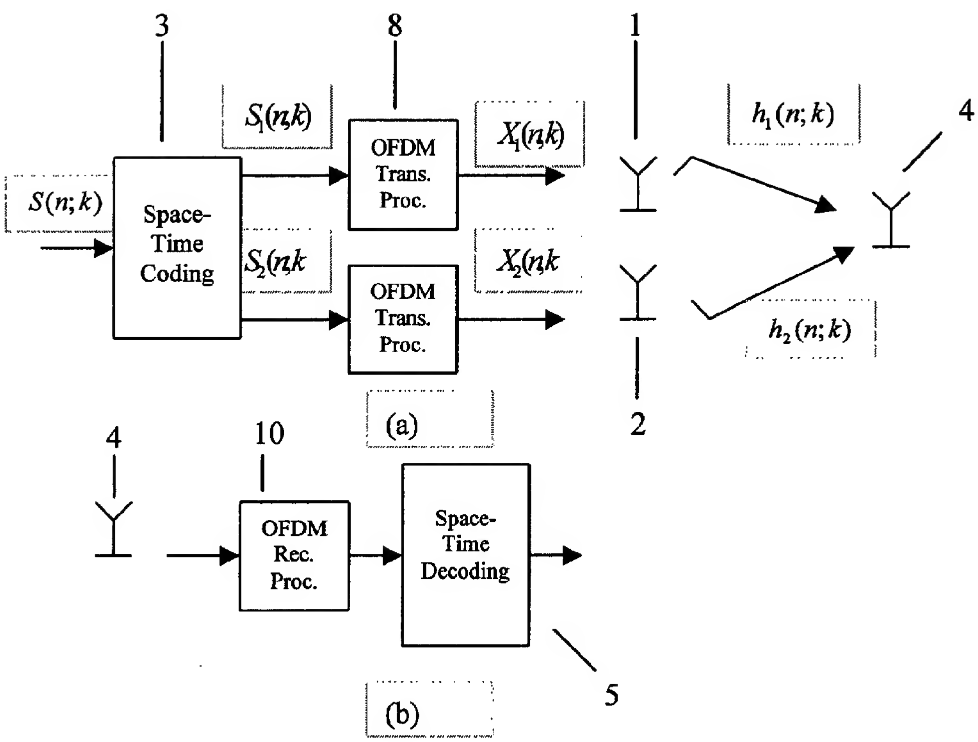

Figure3: Prior Art

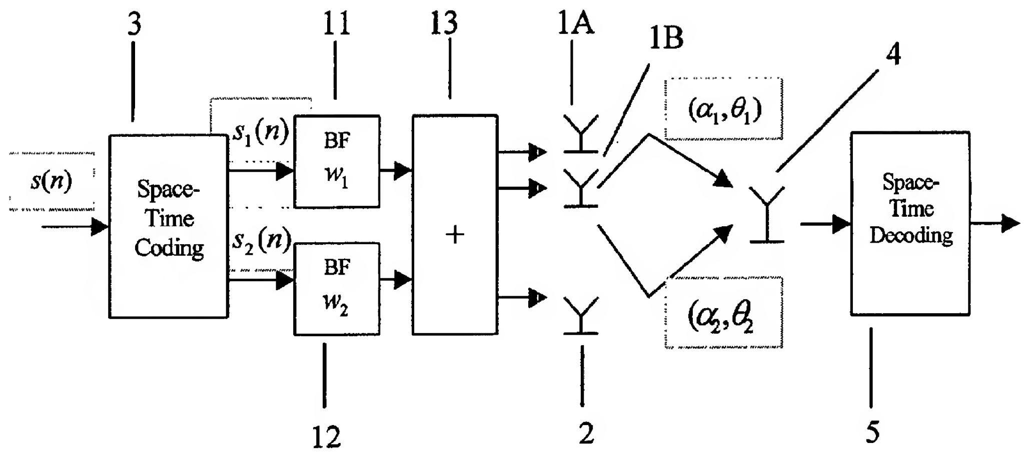

Figure4: Prior Art

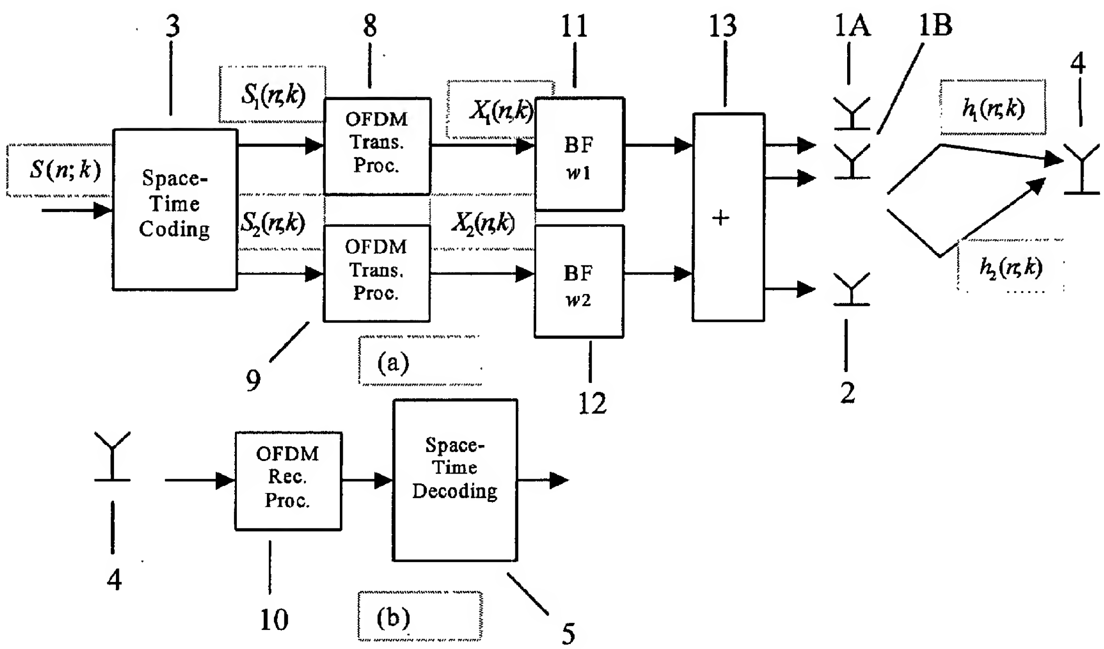
3
8
11
13
1A
1B
4
$S_1(n,k)$
$X_1(n,k)$
$h_1(n,k)$
$S(n;k)$
Space-
Time
Coding
OFDM
Trans.
Proc.
BF
$w1$
+
$S_2(n,k)$
$X_2(n,k)$
$h_2(n,k)$
OFDM
Trans.
Proc.
BF
$w2$
9
(a)
12
2
OFDM
Rec.
Proc.
Space-
Time
Decoding
4
10
(b)
5

Figure5

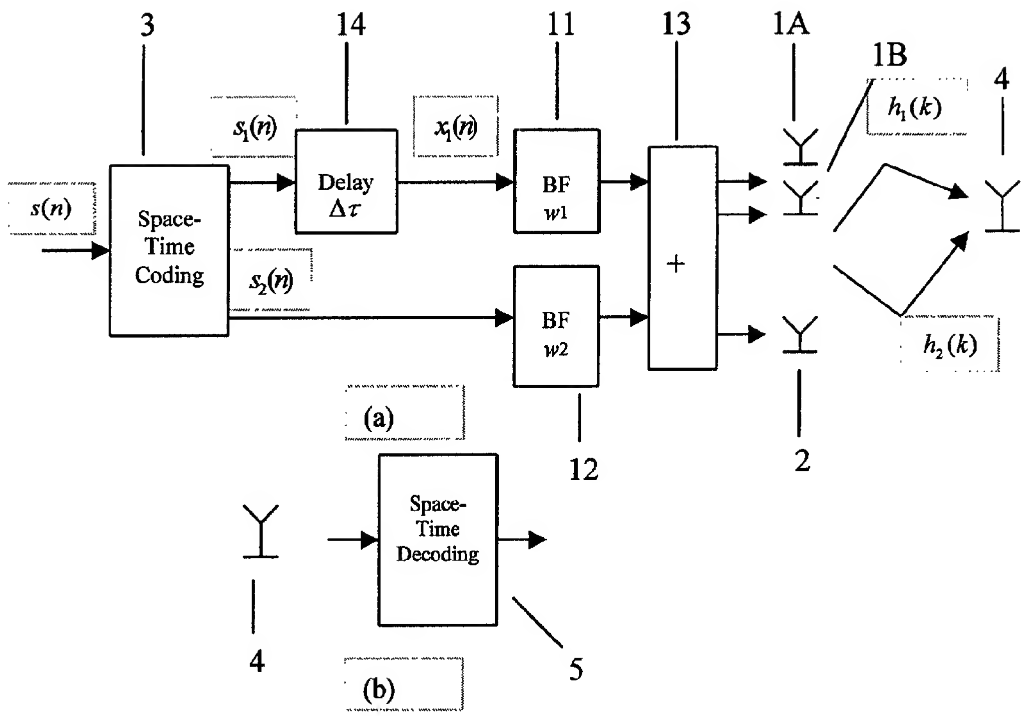
3
14
11
13
1A
1B
4
$s_1(n)$
$x_1(n)$
$h_1(k)$
$s(n)$
Space-Time Coding
Delay $\Delta\tau$
BF $w1$
+
$s_2(n)$
BF $w2$
$h_2(k)$
(a)
12
2
Space-Time Decoding
4
5
(b)

Figure6

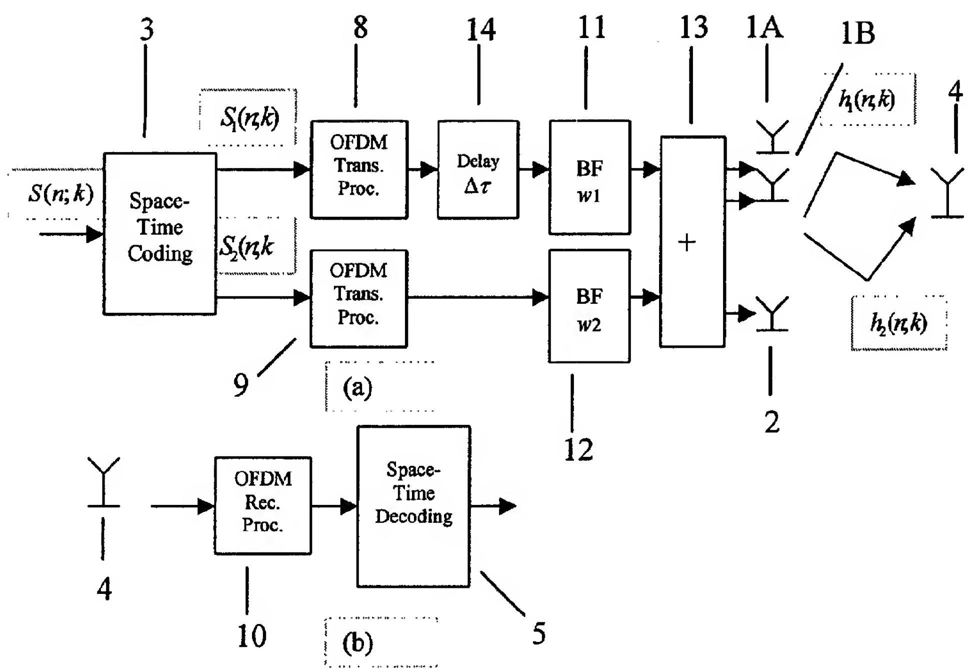
3
8
14
11
13
1A
1B
$S_1(n,k)$
OFDM Trans. Proc.
Delay $\Delta\tau$
BF $w1$
$h_1(n,k)$
4
$S(n;k)$
Space-Time Coding
$S_2(n,k$
+
OFDM Trans. Proc.
BF $w2$
$h_2(n,k)$
9
(a)
2
12
OFDM Rec. Proc.
Space-Time Decoding
4
10
(b)
5

Figure7

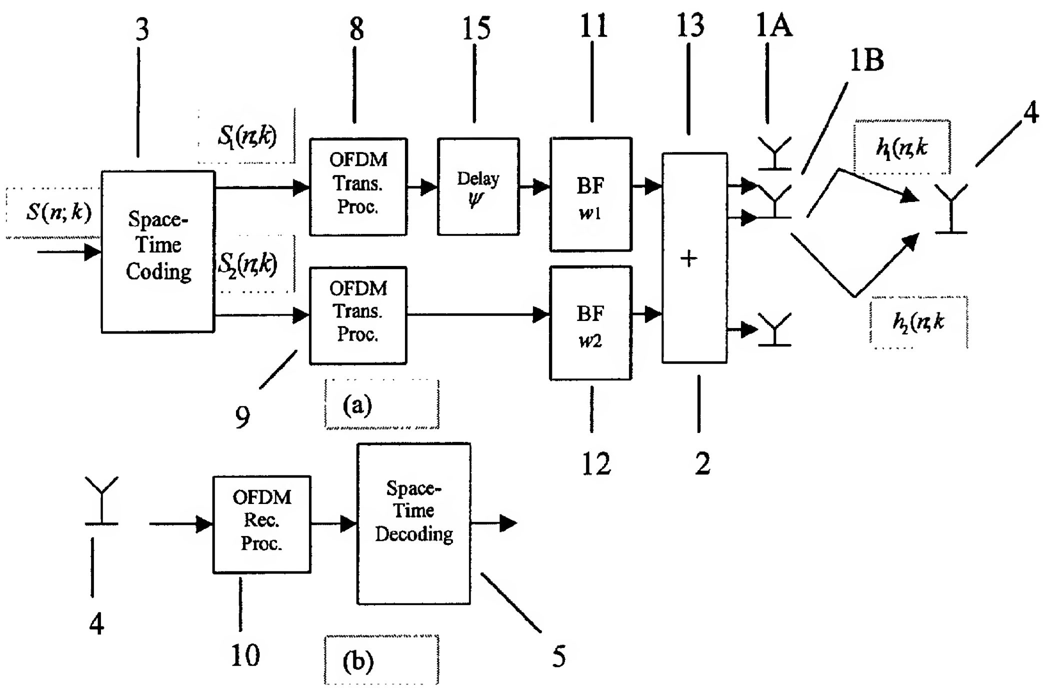
3
8
15
11
13
1A
1B
4
S(n;k)
Space-Time Coding
S1(n,k)
S2(n,k)
OFDM Trans. Proc.
Delay ψ
BF w1
OFDM Trans. Proc.
BF w2
+
h1(n,k
h2(n,k
9
(a)
12
2
OFDM Rec. Proc.
Space-Time Decoding
4
10
(b)
5

Figure8

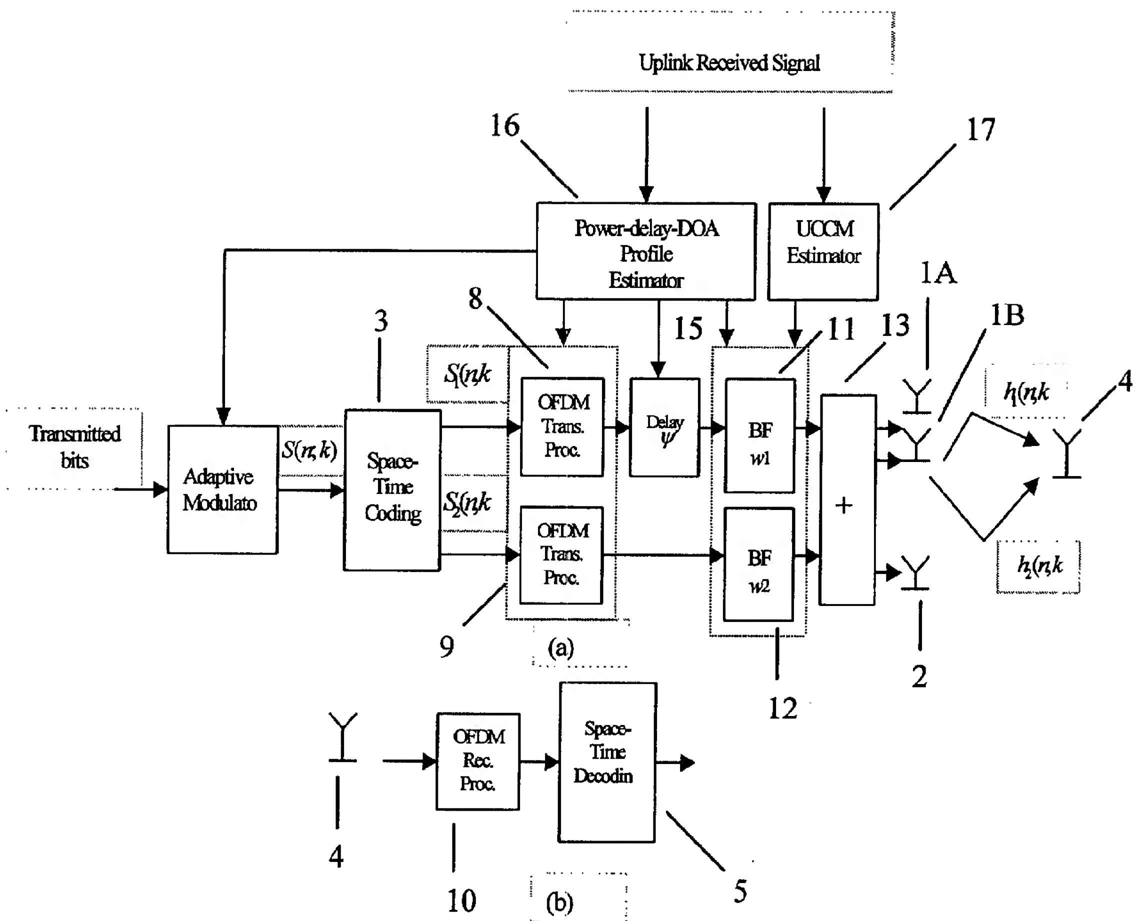
Uplink Received Signal
16
17
Power-delay-DOA Profile Estimator
UCCM Estimator
1A
1B
8
3
15
11
13
$S_1(n,k$
$h_1(n,k$
4
Transmitted bits
Adaptive Modulato
$S(n,k)$
Space-Time Coding
OFDM Trans. Proc.
Delay ψ
BF $w1$
+
$S_2(n,k$
OFDM Trans. Proc.
BF $w2$
$h_2(n,k$
9
(a)
2
12
OFDM Rec. Proc.
Space-Time Decodin
4
10
(b)
5

Figure9

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 02 25 4685

Category	DOCUMENTS CONSIDERED TO BE RELEVANT Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X	SAMPATH H ET AL: "Pre-equalization for MIMO wireless channels with delay spread" VTC 2000-FALL; VEHICULAR TECHNOLOGY CONFERENCE. BOSTON, MA, US, vol. 3 OF 6. CONF. 52, 24 - 28 September 2000, pages 1175-1178, XP010524685 IEEE ISBN: 0-7803-6508-9 * section 1 * * section 2.1 * * sections 3.1, 3.2, 3.3 * ---	1-7	H04L1/06 H04L25/03
X	LIU Z ET AL: "Transmit-antennae space-time block coding for generalized OFDM in the presence on unknown multipath" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. 19, no. 7, July 2001 (2001-07), pages 1352-1364, XP001112325 ISSN: 0733-8716 * abstract * * section II * * figure 1 * ---	1-7	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04L
A	YANG H ET AL: "An adaptive channel precoded space-time transmitter for 3GPP TDD system" GLOBECOM '01. 2001 IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE. SAN ANTONIO, TX, IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE, NEW YORK, NY: IEEE, US, vol. 1 OF 6, 25 - 29 November 2001, pages 529-532, XP001090311 ISBN: 0-7803-7206-9 * section II-B * * figure 3 * --- -/--	1-7	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10 April 2003	Ghigliotti, L

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

EPO FORM 1503 03.82 (P04C01)

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 02 25 4685

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
A,D	ALAMOUTI S M: "A simple transmit diversity technique for wireless communications" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. 16, no. 8, October 1998 (1998-10), pages 1451-1458, XP002100058 ISSN: 0733-8716 * the whole document * ---	1,6	
A,D	TAROKH V ET AL: "Space-time codes for high data rate wireless communication: performance criterion and code construction" IEEE TRANSACTIONS ON INFORMATION THEORY, IEEE INC. NEW YORK, US, vol. 44, no. 2, 1 March 1998 (1998-03-01), pages 744-765, XP002089112 ISSN: 0018-9448 * the whole document * ---	1-7	
			TECHNICAL FIELDS SEARCHED (Int.Cl.7)
X	WO 02 01732 A (NOKIA CORP ;NOKIA INC (US)) 3 January 2002 (2002-01-03) * page 42, line 7 - line 19 * * figures 11,26 * ---	8-45	
X	WO 01 78254 A (NOKIA NETWORKS OY ;KATZ MARCOS (FI); KESKITALO ILKKA (FI); YLITALO) 18 October 2001 (2001-10-18) * page 14, line 10 - page 15, line 14 * * figure 3 * ---	8-45	
	-/--		

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10 April 2003	Ghigliotti, L

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& : member of the same patent family, corresponding document

EPO FORM 1503 03.82 (P04C01)

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 02 25 4685

DOCUMENTS CONSIDERED TO BE RELEVANT			
Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X,D	NEGI R ET AL: "Adaptive antennas for space-time coding over block-time invariant multipath fading channels" VEHICULAR TECHNOLOGY CONFERENCE, 1999 IEEE 49TH HOUSTON, TX, USA 16-20 MAY 1999, PISCATAWAY, NJ, USA,IEEE, US, 16 May 1999 (1999-05-16), pages 70-74, XP010341916 ISBN: 0-7803-5565-2 * section 3 *	8-22, 24-28	
A,D	EP 1 087 545 A (UNIV SINGAPORE) 28 March 2001 (2001-03-28) * the whole document *	8-45	
			TECHNICAL FIELDS SEARCHED (Int.Cl.7)

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10 April 2003	Ghigliotti, L

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D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

European Patent Office

Application Number

EP 02 25 4685

CLAIMS INCURRING FEES

The present European patent application comprised at the time of filing more than ten claims.

- [ ] Only part of the claims have been paid within the prescribed time limit. The present European search report has been drawn up for the first ten claims and for those claims for which claims fees have been paid, namely claim(s):

- [ ] No claims fees have been paid within the prescribed time limit. The present European search report has been drawn up for the first ten claims.

LACK OF UNITY OF INVENTION

The Search Division considers that the present European patent application does not comply with the requirements of unity of invention and relates to several inventions or groups of inventions, namely:

see sheet B

- [x] All further search fees have been paid within the fixed time limit. The present European search report has been drawn up for all claims.

- [ ] As all searchable claims could be searched without effort justifying an additional fee, the Search Division did not invite payment of any additional fee.

- [ ] Only part of the further search fees have been paid within the fixed time limit. The present European search report has been drawn up for those parts of the European patent application which relate to the inventions in respect of which search fees have been paid, namely claims:

- [ ] None of the further search fees have been paid within the fixed time limit. The present European search report has been drawn up for those parts of the European patent application which relate to the invention first mentioned in the claims, namely claims:

European Patent Office

LACK OF UNITY OF INVENTION SHEET B

Application Number
EP 02 25 4685

The Search Division considers that the present European patent application does not comply with the requirements of unity of invention and relates to several inventions or groups of inventions, namely:

1. Claims: 1-7

   Method and system for achieving diversity gain in frequency selective fading channel, by means of space-time encoding and zero-forcing pre-equalisation.

2. Claims: 8-45

   Method and system for achieving simultaneous beamforming and transmit diversity gain by means of a a space-time encoder and a transmit beamformer.

ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 02 25 4685

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report.
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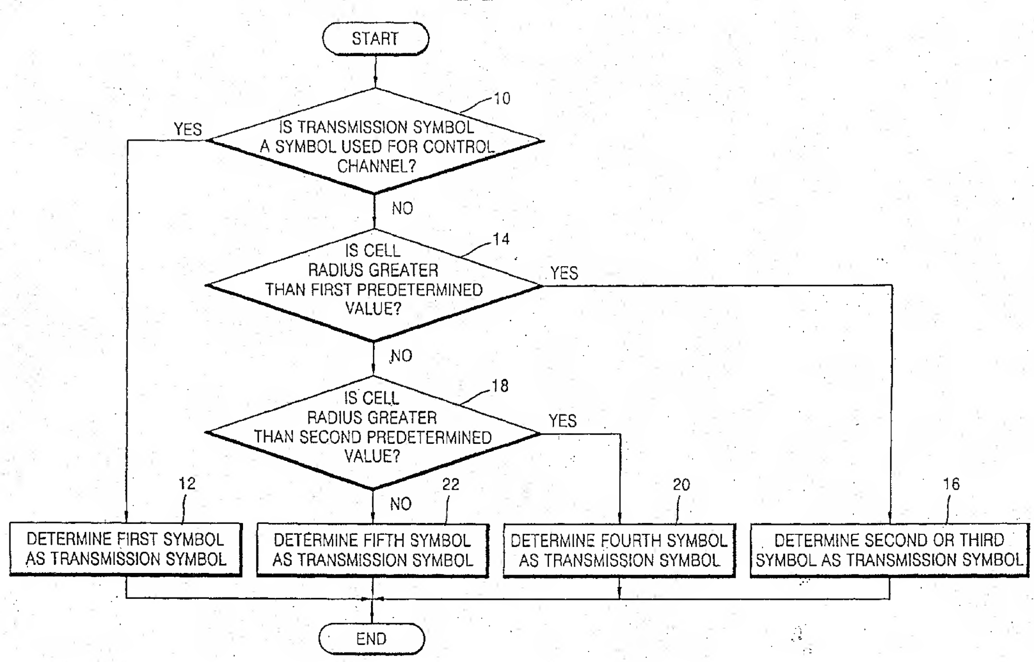
(54) Multicarrier transmission with adaptation to channel characteristics

(57) An orthogonal frequency division multiplexing (OFDM) communication method and apparatus adapted to channel characteristics are provided. The OFDM communication method includes changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol depending on a type of the transmission symbol and a radius of a cell, in which communication is performed. The OFDM communication apparatus includes a symbol inspector, for inspecting a type of a transmission symbol and outputting the result of the inspection as a first control signal, and a symbol and format converter, for changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol in response to the first control signal and a radius of a cell, in which communication is performed.

FIG. 1



EP 1 387 545 A2

Description

[0001] The present invention relates to orthogonal frequency division multiplexing (OFDM) communication, and more particularly, to an OFDM communication method and apparatus adapted to channel characteristics.

[0002] With a variety of environments in which a communication method is used, the communication method is required to be effective even if Doppler frequency or delay spread changes. However, since an optimum physical layer varies with channel change speed and delay spread, it is difficult to efficiently support a communication method using a single physical layer. Accordingly, a hierarchical cell including a variety of cells is used in a single communication method.

[0003] When using such a hierarchical cell, channels for users corresponding to different layers have different characteristics. For example, when a cell has a large radius, delay spread is long, and a channel change speed is fast. Accordingly, if the same modulation method is applied to different layers, a communication method cannot be adapted to the channel characteristics. In order to overcome this problem, a conventional communication method uses OFDM when the channel change speed is slow and uses code division multiple access (CDMA) when the channel change speed is fast. As described above, when using the conventional communication method, two modems of different types need to be provided for a terminal. Accordingly, the conventional communication method increases the complexity of transmitter and receiver of a terminal. In addition, since signals having different spectrum characteristics are used, the conventional communication method is difficult to develop, and radio resource management such as handover and association is difficult.

[0004] According to an aspect of the present invention, there is provided an OFDM communication method adapted to channel characteristics, including changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol depending on a type of the transmission symbol and a radius of a cell, in which communication is performed.

[0005] The invention thus provides an orthogonal frequency division multiplexing (OFDM) communication method through which at least one of the length of a transmission symbol, the format of a transmission symbol, and the format of a frame is changed to adapt to channel characteristics such as channel change speed and channel spread.

[0006] According to another aspect of the present invention, there is provided an OFDM communication apparatus adapted to channel characteristics, including a symbol inspector, which inspects a type of a transmission symbol and outputs the result of the inspection as a first control signal; and a symbol and format converter, which changes at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol in response to the first control signal and a radius of a cell, in which communication is performed.

[0007] The invention thus also provides an OFDM communication apparatus for performing the OFDM communication method of the invention, which is adapted to the channel characteristics.

[0008] The above and other features and advantages of the present invention will become more apparent by describing in detail preferred embodiments thereof with reference to the attached drawings in which:

FIG. 1 is a flowchart of an orthogonal frequency division multiplexing (OFDM) communication method adapted to channel characteristics according to a first embodiment of the present invention;
FIG. 2 is a diagram showing an example of a single frame including symbols having various lengths;
FIG. 3 is a flowchart of an OFDM communication method adapted to channel characteristics according to a second embodiment of the present invention;
FIG. 4 is a diagram showing an example of a macro format;
FIG. 5 is a diagram showing an example of a micro format;
FIG. 6 is a diagram showing an example of a pico format;
FIG. 7 is a flowchart of an embodiment of step 16 shown in FIG. 1 according to the present invention;
FIG. 8 is a diagram showing a hierarchical cell structure;
FIG. 9 is a diagram showing an example of a usual multiplex carrier wave transmission symbol;
FIG. 10 is a diagram showing another example of a usual multiplex carrier wave transmission symbol;
FIG. 11 is a diagram showing still another example of a usual multiplex carrier wave transmission symbol;
FIG. 12 is a block diagram of an OFDM communication apparatus for performing an OFDM communication method of the present invention, according to an embodiment of the present invention;
FIG. 13 is a block diagram of an embodiment of a symbol and format converter shown in FIG. 12;
FIG. 14 is a block diagram of another embodiment of the symbol and format converter shown in FIG. 12;
FIG. 15 is a block diagram of a first converter shown in FIG. 13;
FIG. 16 is a graph showing changes in a bit error rate with respect to changes in Doppler frequency; and
FIG. 17 is a graph showing changes in a bit error rate with respect to changes in the number of carrier waves.

[0009] Hereinafter, preferred embodiments of an orthogonal frequency division multiplexing (OFDM) communication

method to adapt to channel characteristics according to the present invention will be described in detail with reference to the attached drawings. In an OFDM communication method to adapt to channel characteristics according to the present invention, at least one of the length of a transmission symbol, the format of a frame, and the format of a transmission symbol is changed depending on a type of transmission symbol and the radius of a cell, in which communication is performed.

[0010] Channel variation is usually measured in terms of Doppler frequency multiplied by the length of an OFDM symbol, denoted as fdTs (fd: Doppler frequency in Hz, Ts; symbol duration in seconds). When fdTs is less than 0.01, the effect of channel variation on the detection performance is negligible. However, when fdTs becomes greater than 0.01, the effect becomes noticeable. Generally speaking, a fast channel change speed is when fdTs is greater than 0.01, though this is not a hard and fast rule.

[0011] Likewise, the channel length is measured by the delay spread of a channel, which is the time delay incurred from when the first signal components arrive at the receiver to when the last signal components arrive at the receiver. For example, if the last signal arrives at the receiver 0.01 seconds after the first signal arrived at the receiver, the length of the channel is 0.01 seconds. A long, medium or short channel is a relative measure utilized in the industry to describe this relative channel length. For example, if the default symbol length is 0.1 msec, a 0.1-msec channel is considered to be a long channel, a 0.01-msec channel is considered to be a medium channel, and a 0.001-msec is considered to be a short channel. In other words, when the length of a channel, divided by the length of the default OFDM symbol, is more than 10%, it is considered to be a long channel.

[0012] FIG. 1 is a flowchart of an OFDM communication method to adapt to channel characteristics according to a first embodiment of the present invention. The OFDM communication method according to the first embodiment includes determining the length of a transmission symbol depending on a type of transmission symbol and a cell radius (steps 10 through 22).

[0013] FIG. 2 is a diagram showing an example of a single frame 40, in which symbols having various lengths are mixed. The single frame 40 includes first symbols 42 and 44, second symbols 50 and 52, third symbols 54, 56, 58, and 60, a fourth symbol 48, and a fifth symbol 46.

[0014] In the OFDM communication method according to the first embodiment of the present invention shown in FIG. 1, the length of a transmission symbol is changed depending on a type of transmission symbol and the radius of a cell, in which communication is performed.

[0015] More specifically, it is determined whether a transmission symbol is a symbol that is used for a control channel in step 10. If it is determined that the transmission symbol is the symbol that is used for the control channel, the first symbol 42 or 44 shown in FIG. 2 is determined as the transmission symbol in step 12. The first symbol 42 or 44 contains control information and has a length A. In other words, if it is determined that the transmission symbol is the symbol that is used for the control channel, the length of the transmission symbol is set to A. As described above, when a large amount of data is not necessary or when it is necessary to finely divide time, as in random access or control, the relatively short length A is determined as the length of a transmission symbol.

[0016] If it is determined that the transmission symbol is not the symbol that is used for the control channel, it is determined whether a cell radius is greater than a first predetermined value in step 14. If it is determined that the cell radius is greater than the first predetermined value, the second symbol 50 or 52 or the third symbol 54, 56, 58, or 60 shown in FIG. 2 is determined as the transmission symbol in step 16. The second symbol 50 or 52 has a length B and is suitable to channel characteristics, in which a channel change speed is slow and the length of a channel is long. The third symbol 54, 56, 58, or 60 has a length C and is suitable to channel characteristics, in which a channel change speed is fast and the length of a channel is long or short. In other words, if it is determined that the cell radius is greater than the first predetermined value, the length of the transmission symbol is set to B or C.

[0017] However, if it is determined that the cell radius is not greater than the first predetermined value, it is determined whether the cell radius is greater than a second predetermined value in step 18. Here, the second predetermined value is less than the first predetermined value. If it is determined that the cell radius is greater than the second predetermined value, the fourth symbol 48 shown in FIG. 2 is determined as the transmission symbol in step 20. The fourth symbol 48 has a length D and is suitable to channel characteristics, in which a channel change speed and the length of a channel are medium. In other words, if it is determined that the cell radius is not greater than the first predetermined value but greater than the second predetermined value, the length of the transmission symbol is set to D.

[0018] However, if it is determined that the cell radius is not greater than the second predetermined value, the fifth symbol 46 shown in FIG. 2 is determined as the transmission symbol in step 22. The fifth symbol 46 has a length E and is suitable to channel characteristics, in which a channel change speed is slow and the length of a channel is short. In other words, if it is determined that the cell radius is not greater than the second predetermined value, the length of the transmission symbol is set to E.

[0019] According to the present invention, the length D of the fourth symbol 48 is shorter than the length B of the second symbol 50, and each of the lengths A, C, and E of the respective first, third, and fifth symbols 42, 54, and 46 is shorter than the length D of the fourth symbol 48. In addition, according to the present invention, each of the lengths

B, C, D, and E of the respective second, third, fourth, and fifth symbols 50, 54, 48, and 46 may be an integer multiple of the length A of the first symbol 42, and each of the lengths B, C, and D of the respective second, third, and fourth symbols 50, 54, and 48 may be an integer multiple of the length E of the fifth symbol 46.

[0020] In order to change the length of a transmission symbol, as shown in FIG. 1, the present invention changes the number of carrier waves while fixing an entire signal bandwidth. The entire signal bandwidth indicates the result of dividing an interval between carrier waves by the length of a transmission symbol. For example, when increasing the number of carrier waves while fixing an entire signal bandwidth, a distance between carrier waves is long and the length of a transmission symbol increases. Conversely, when decreasing the number of carrier waves while fixing an entire signal bandwidth, a distance between carrier waves is short and the length of a transmission symbol decreases. As described above, the length of a transmission symbol can be changed by adjusting the number of carrier waves, in step 12, 16, 20 or 22.

[0021] FIG. 3 is a flowchart of an OFDM communication method to adapt to channel characteristics according to a second embodiment of the present invention. The OFDM communication method includes converting the format of a frame depending on a cell radius in steps 70 through 78.

[0022] FIG. 4 is a diagram showing an example of a macro format. A single frame 90 is composed of a single first symbol and a plurality of second symbols, and a plurality of third symbols.

[0023] FIG. 5 is a diagram showing an example of a micro format. A single frame 92 is composed of a single first symbol and a plurality of fourth symbols.

[0024] FIG. 6 is a diagram showing an example of a pico format. A single frame 94 is composed of a single first symbol and a plurality of fifth symbols.

[0025] In the OFDM communication method according to the second embodiment of the present invention shown in FIG. 3, the format of a frame is converted depending on the radius of a cell, in which communication is performed.

[0026] For this operation, it is determined weather a cell radius is greater than a first predetermined value in step 70. If it is determined that the cell radius is greater than the first predetermined value, the format of a frame is converted into a macro format, as shown in FIG. 4, in step 72. Referring to FIG. 4, the macro format is composed of a single first symbol, a plurality of second symbols, and a plurality of third symbols. In other words, when a channel change speed is fast or slow and the length of a channel is long due to a large cell radius, the format of the frame is converted into the macro format shown in FIG. 4.

[0027] However, if it is determined that the cell radius is not greater than the first predetermined value, it is determined weather the cell radius is greater than a second predetermined value in step 74. The second predetermined value is smaller than the first predetermined value. If it is determined that the cell radius is greater than the second predetermined value, the format of a frame is converted into a micro format, as shown in FIG. 5, in step 76. Referring to FIG. 5, the micro format is composed of a single first symbol and a plurality of fourth symbols. In other words, when a channel change speed and the length of a channel are medium, the format of the frame is converted into the micro format.

[0028] However, if it is determined that the cell radius is not greater than the second predetermined value, the format of a frame is converted into a pico format, as shown in FIG. 6, in step 78. Referring to FIG. 6, the pico format is composed of a single first symbol and a plurality of fifth symbols. In other words, when a channel change speed is slow and the length of a channel is short due to a small cell radius, the format of the frame is converted into the pico format.

[0029] FIG. 7 is a flowchart of an embodiment of step 16 shown in FIG. 1 according to the present invention. The embodiment of step 16 includes determining a second or third symbol as a transmission symbol depending on a channel change speed in steps 110 through 114.

[0030] Referring to FIG. 7, if it is determined that the cell radius is greater than the first predetermined value (step 14 of FIG. 1), it is determined whether a channel change speed is greater than a predetermined speed in step 110.

[0031] If it is determined that the channel change speed is not greater than the predetermined speed, the second symbol 50 shown in FIG. 2 is determined as the transmission symbol in step 112. In other words, the length of the transmission symbol is set to B. However, if it is determined that the channel change speed is greater than the predetermined speed, the third symbol 54 is determined as the transmission symbol in step 114. In other words, the length of the transmission symbol is set to C.

[0032] According to a third embodiment of the present invention, the length of a transmission symbol and the format of a frame are changed depending on a type of transmission symbol and a cell radius. For this operation, referring to FIG. 1, if it is determined that the cell radius is greater than the first predetermined value, the second or third symbol is determined as the transmission symbol, and simultaneously the format of the frame is converted into the macro format shown in FIG. 4, in step 16. However, if it is determined that the cell radius is not greater than the first predetermined value but is greater than the second predetermined value, the fourth symbol is determined as the transmission symbol, and simultaneously the format of the frame is converted into the micro format shown in FIG. 5, in step 20. In addition, if it is determined that the cell radius is not greater than the second predetermined value, the fifth symbol is determined as the transmission symbol, and simultaneously the format of the frame is converted into the pico format shown in FIG. 6, in step 22.

[0033] FIG. 8 is a diagram showing a hierarchical cell structure, which is composed of macro cells 130, micro cells 132, and pico cells 134.

[0034] Referring to FIG. 8, the macro cells 130 represented by dotted lines correspond to cells having a radius that is greater than the first predetermined value. The micro cells 132 represented by bold sold lines correspond to cells having a radius that is not greater than the first predetermined value but greater than the second predetermined value. The pico cells 134 represented by thin solid lines correspond to cells having a radius that is not greater than the second predetermined value. The hierarchical cell structure shown in FIG. 8 is used in order to increase frequency efficiency when frequency resources are limited. As shown in FIG. 8, a plurality of micro cells 132 exist within each macro cell 130, and a plurality of pico cells 134 exist within each micro cell 132. Usually, the hierarchical cell structure is designed such that users with a fast channel change speed are gathered at the macro cells 130 and users with a slow channel change speed are gathered in the micro cells 132 or the pico cells 134. This is disclosed in pages 301-304 of a book entitled "Radio Resource Management for Wireless Networks", written by Jens Zander and Seong-Lyun Kim, and published by Artech Houser in 2001.

[0035] FIG.s 9 to 11 show examples of the multiplex carrier wave transmission symbol. FIG. 9 is an example of the second symbol and FIG.s 10 and 11 are different examples of the third symbol. In the following description, the different parts of the symbols of different examples are each given different names to avoid confusion. Therefore, the existence of a "third cyclic prefix" (for example) in a symbol should not be understood as requiring a "first" or "second" cyclic prefix in that symbol.

[0036] FIG. 9 is a diagram showing an example of a multiplex carrier wave transmission symbol. In this example, the transmission symbol is composed of a first cyclic prefix (CP) 150, a first transmission signal 158, and a first cyclic suffix (CS) 154.

[0037] According to the fourth embodiment of the present invention, the format of a symbol as well as the length of the symbol can be changed depending on a cell radius and a channel change speed.

[0038] For example, if it is determined that the channel change speed is not greater than the predetermined speed, the second symbol is determined as the transmission symbol and the format of the second symbol is converted into a format shown in FIG. 9 in step 112 of FIG. 7. In FIG. 9, the first CP 150 of the transmission symbol is the result of copying an end portion 152 of the first transmission signal 158 to the front of the first transmission signal 158 and is used to eliminate the interference of a previous symbol. The first CS 154 of the transmission symbol is the result of copying a beginning portion 156 of the first transmission signal 158 to the back of the first transmission signal 158 and is used to mitigate the alignment condition of transmission time when a carrier wave is divided and used by multiple users usually in an upward channel. Here, the first transmission signal 158 contains transmission data. As described above, since the end portion 152 of the transmission data 158 is copied to the first CP 150 and the beginning portion 156 of the transmission data 158 is copied to the first CS 154, the transmission symbol shown in FIG. 9 has a cyclic structure.

[0039] FIG. 10 is a diagram showing another example of a multiplex carrier wave transmission symbol. In this example, the transmission symbol is composed of a second CP 170, second and third transmission signals 172 and 174, and a second CS 176.

[0040] FIG. 11 is a diagram showing still another example of a multiplex carrier wave transmission symbol. In this example, the transmission symbol is composed of a third CP 190, fourth and fifth transmission signals 192 and 194, and a third CS 196.

[0041] However, if it is determined that the channel change speed is greater than the predetermined speed, the third symbol is determined as the transmission symbol and the format of the third symbol is converted into a format shown in FIG. 10 or 11 in step 114 of FIG. 7.

[0042] According to the present invention, the second CP 170 of the third symbol shown in FIG. 10 includes the end portion of transmission data stored in each of the second and third transmission signals 172 and 174 and the beginning portion of the transmission data. In other words, the second CP 170 is composed of two first CPs 150 and one first CS 154 shown in FIG. 9. In addition, each of the second and third transmission signals 172 and 174 shown in FIG. 10 contains the same transmission data as that contained in the first transmission signal 158 shown in FIG. 9. Unlike the transmission symbol shown in FIG. 9, the transmission symbol shown in FIG. 10 includes repeated transmission data following the second CP 170. Here, the second CS 176 includes the beginning portion of the transmission data. In other words, the second CS 176 is composed of one first CS 154 shown in FIG. 9.

[0043] According to the present invention, the third CP 190 of the third symbol shown in FIG. 11 includes the end portions of transmission data stored in each of the fourth and fifth transmission signals 192 and 194. In other words, the third CP 190 is composed of only two first CPs 150. In addition, each of the fourth and fifth transmission signals 192 and 194 shown in FIG. 11 contains the same transmission data as that contained in the first transmission signal 158 shown in FIG. 9. Unlike the transmission symbol shown in FIG. 9, the transmission symbol shown in FIG. 11 includes repeated transmission data following the third CP 190. Here, the third CS 196 includes the beginning portions of the transmission data. In other words, the third CS 196 is composed of two first CSs 154 shown in FIG. 9. The

transmission symbol shown in FIG. 11 can be used when timing does not agree well as a whole as in random access.

[0044] Consequently, in an OFDM communication method, the length of the first CP 150 is required to be longer than a channel length. However, since duplicate information is contained in the first CP 150, communication efficiency is decreased when the result of dividing the length of the first CP 150 by the transmission data contained in the first transmission signal 158 is too large. Accordingly, in order to maintain the communication efficiency, the length of the transmission data needs to be 5-10 times longer than the length of the first CP 150. In other words, in the transmission symbol, the length of the first CP 150 needs to be as short as possible. Here, if the length of the transmission symbol is long in a state in which a channel change speed is fast, a channel may change within the transmission data, and thus communication performance may be degraded. As described above, it is necessary to increase the length of the first CP 150 as a channel length increases, but there is a limitation in increasing the length of the first CP 150. In order to solve this problem, an OFDM communication method according to the present invention described above adaptively changes the length of a transmission symbol depending on a channel change speed and a cell radius, as shown in FIG. 1. As described above, influence of inter symbol interference (ISI) can be overcome while the degree of overhead due to the first CP 150 is fixed, by adaptively changing the length of a transmission symbol.

[0045] Hereinafter, the structure and operation of an OFDM communication apparatus adapted to channel characteristics according to the present invention, which performs the above-described OFDM communication method adapted to channel characteristics according to the present invention, will be described with reference to the attached drawings.

[0046] FIG. 12 is a block diagram of an OFDM communication apparatus for performing the above-described OFDM communication method according to an embodiment of the present invention. The OFDM communication apparatus includes a symbol inspector 210 and a symbol and format converter 212.

[0047] Referring to FIG. 12, in order to perform step 10 shown in FIG. 1, the symbol inspector 210 inspects the type of a transmission symbol that is input through an input terminal IN1 and outputs the result of the inspection to the symbol and format converter 212 as a first control signal C1. For example, the symbol inspector 210 inspects whether the transmission symbol input through the input terminal IN1 is a symbol used for a control channel and outputs the result of the inspection as the first control signal C1.

[0048] In order to perform steps 12 through 22 shown in FIG. 1, the symbol and format converter 212 changes at least one of the length of the transmission symbol, the format of a frame, and the format of the transmission symbol in response to a cell radius that is input through an input terminal IN2 and the first control signal C1 received from the symbol inspector 210, and outputs the result of the change through an output terminal OUT1. Here, what will be changed among the length of the transmission symbol, the format of a frame, and the format of the transmission symbol is predetermined.

[0049] The following description concerns the structures and operations of embodiments of the symbol and format converter 212 shown in FIG. 12 according to the present invention. FIG.s 13 and 14 each show an example of the converter 212. In the following description, the different parts of the converter in the two examples are each given different names to avoid confusion. Therefore, the existence of a "third comparator" (for example) in the converter should not be understood as requiring a "first" or "second" comparator in that converter.

[0050] FIG. 13 is a block diagram of an embodiment 212A of the symbol and format converter 212 shown in FIG. 12. The embodiment 212A includes a first comparator 230, a second comparator 232, and a first converter 234.

[0051] In order to perform step 14 shown in FIG. 1, the first comparator 230 of the symbol and format converter 212A shown in FIG. 13 compares the cell radius that is input through an input terminal IN3 with a first predetermined value in response to the first control signal C1 that is input from the symbol inspector 210 and outputs the result of the comparison to the second comparator 232 and the first converter 234 as a second control signal C2. In other words, when it is recognized based on the first control signal C1 received from the symbol inspector 210 that the transmission symbol is not a symbol used for a control channel, the first comparator 230 compares the cell radius with the first predetermined value. Here, the first predetermined value may be set in the first comparator 230 in advance, as shown in FIG. 13, or may be externally input, unlike the structure shown in FIG. 13.

[0052] In order to perform step 18 shown in FIG. 1, the second comparator 232 compares the cell radius with a second predetermined value in response to the second control signal C2 received from the first comparator 230 and outputs the result of the comparison to the first converter 234 as a third control signal C3. For example, when it is recognized based on the second control signal C2 received from the first comparator 230 that the cell radius is not greater than the first predetermined value, the second comparator 232 compares the cell radius with the second predetermined value and outputs the result of the comparison as the third control signal C3. Here, the second predetermined value may be set in the second comparator 232 in advance, as shown in FIG. 13, or may be externally input, unlike the structure shown in FIG. 13.

[0053] In order to perform steps 12, 16, 20, and 22 shown in FIG. 1, the first converter 234 determines one among first through fifth symbols as the transmission symbol in response to the first control signal C1 received from the symbol inspector 210, the second control signal C2 received from the first comparator 230, and the third control signal C3

received from the second comparator 232 and outputs the determined symbol through an output terminal OUT2. For example, in order to perform step 12, when it is recognized based on the first control signal C1 received from the symbol inspector 210 that the transmission symbol is a symbol used for a control channel, the first converter 234 determines the length of the transmission symbol as A. However, in order to perform step 16, when it is recognized based on the first and second control signals C1 and C2 that the transmission symbol is not a symbol used for a control channel and the cell radius is greater than the first predetermined value, the first converter 234 determines the length of the transmission symbol as B or C. In addition, in order to perform step 20, when it is recognized based on the first, second, and third control signals C1, C2 and C3 that the transmission symbol is not a symbol used for a control channel and the cell radius is not greater than the first predetermined value but is greater than the second predetermined value, the first converter 234 determines the length of the transmission symbol as D. In order to perform step 22, when it is recognized based on the first, second, and third control signals C1, C2 and C3 that the transmission symbol is not a symbol used for a control channel and the cell radius is not greater than the first predetermined value and is not greater than the second predetermined value, the first converter 234 determines the length of the transmission symbol as E.

[0054] FIG. 14 is a block diagram of another embodiment 212B of the symbol and format converter 212 shown in FIG. 12. The embodiment 212B includes a third comparator 250, a fourth comparator 252, and a second converter 254.

[0055] The symbol and format converter 212B shown in FIG. 14 performs the OFDM communication method shown in FIG. 3. In order to perform step 70 shown in FIG. 3, the third comparator 250 of the symbol and format converter 212B shown in FIG. 14 compares the cell radius that is input through an input terminal IN4 with the first predetermined value and outputs the result of the comparison to the fourth comparator 252 and the second converter 254 as a fourth control signal C4. Here, the first predetermined value may be set in the third comparator 250 in advance, as shown in FIG. 14, or may be externally input, unlike the structure shown in FIG. 14.

[0056] In order to perform step 74 shown in FIG. 3, the fourth comparator 252 compares the cell radius input through the input terminal IN4 with the second predetermined value in response to the fourth control signal C4 received from the third comparator 250 and outputs the result of the comparison to the second converter 254 as a fifth control signal C5. For example, when it is recognized based on the fourth control signal C4 received from the third comparator 250 that the cell radius is not grater than the first predetermined value, the fourth comparator 252 compares the cell radius with the second predetermined value and outputs the result of the comparison as the fifth control signal C5.

[0057] In order to perform steps 72, 76, and 78 of FIG. 3, the second converter 254 converts the format of a frame into a macro format, micro format, or pico format in response to the fourth control signal C4 received from the third comparator 250 and the fifth control signal C5 received from the fourth comparator 252 and outputs the frame having the converted format through an output terminal OUT3. For example, in order to perform step 72, when it is recognized based on the fourth control signal C4 that the cell radius is grater than the first predetermined value, the second converter 254 converts the format of a frame into the macro format shown in FIG. 4. In order to perform step 76, when it is recognized based on the fourth and fifth control signals C4 and C5 that the cell radius is not grater than the first predetermined value but is greater than the second predetermined value, the second converter 254 converts the format of a frame into the micro format shown in FIG. 5. In order to perform step 78, when it is recognized based on the fourth and fifth control signals C4 and C5 that the cell radius is less than both first and second predetermined values, the second converter 254 converts the format of a frame into the pico format shown in FIG. 6.

[0058] According to an embodiment of the present invention, the symbol and format converter 212 shown in FIG. 12 may be provided with the symbol and format converter 212A shown in FIG. 13 in order to perform steps 12 through 22 shown in FIG. 1 and the symbol and format converter 212B shown in FIG. 14 in order to perform the OFDM communication method shown in FIG. 3.

[0059] According to another embodiment of the present invention, the symbol and format converter 212 shown in FIG. 12 may be provided with only the symbol and format converter 212A shown in FIG. 13 in order to perform steps 12 through 22 shown in FIG. 1 and the OFDM communication method shown in FIG. 3. In this situation, the symbol and format converter 212A shown in FIG. 13 can perform all of the steps 12 through 22 shown in FIG. 1 and the OFDM communication method shown in FIG. 3. For example, the first and second comparators 230 and 232 perform steps 14 and 18, respectively, shown in FIG. 1 and also perform steps 70 and 74, respectively, shown in FIG. 3. The first converter 234 performs steps 12, 16, 20, 22, 72, 76, and 78. In other words, the first converter 234 converts the format of a frame into a macro, micro, or pico format in response to the second and third control signals C2 and C3 respectively received from the first and second comparators 230 and 232 and outputs the frame having the converted format through the output terminal OUT2. For example, in order to perform step 72, when it is recognized based on the second control signal C2 that the cell radius is greater than the first predetermined value, the first converter 234 converts the format of a frame into the macro format shown in FIG. 4. In order to perform step 76, when it is recognized based on the second and third control signals C2 and C3 that the cell radius is not greater than the first predetermined value but is greater than the second predetermined value, the first converter 234 converts the format of a frame into the micro format shown in FIG. 5. In order to perform step 78, when it is recognized based on the second and third control signals C2 and C3 that the cell radius is not greater than both first and second predetermined values, the first converter 234

converts the format of a frame into the pico format shown in FIG. 6.

[0060] FIG. 15 is a block diagram of the first converter 234 shown in FIG. 13. The first converter 234 includes a fifth comparator 270 and a format converter 272.

[0061] The first converter 234 shown in FIG. 13 may include the fifth comparator 270 in order to perform step 110 shown in FIG. 7. In this situation, the fifth comparator 270 compares a channel change speed with a predetermined speed in response to the second control signal C2 received from the first comparator 230 and outputs the result of the comparison as a sixth control signal C6. For example, when it is recognized based on the second control signal C2 that the cell radius is greater than the first predetermined value, the fifth comparator 270 compares the channel change speed with the predetermined speed and outputs the result of the comparison as the sixth control signal C6. Here, the format converter 272 of the first converter 234 converts the format of the determined transmission symbol in response to the sixth control signal C6 received from the fifth comparator 270 and outputs the transmission symbol having the converted format through the output terminal OUT2. For example, when it is recognized based on the sixth control signal C6 received from the fifth comparator 270 that the channel change speed is not greater than the predetermined speed, the format converter 272 converts the format of the transmission symbol into the format shown in FIG. 9 in order to perform step 112. However, when it is recognized based on the sixth control signal C6 received from the fifth comparator 270 that the channel change speed is greater than the predetermined speed, the format converter 272 converts the format of the transmission symbol into the format shown in FIG. 10 or 11 in order to perform step 114.

[0062] When an OFDM communication method and apparatus adapted to channel characteristics according to the present invention are used at a whole signal bandwidth of 20 MHz, the results of operation are obtained, as shown in Table 1.

Table 1

Division	First symbol	Second symbol	Third symbol	Fourth symbol	Fifth symbol
Number of carrier waves	512	4096	1024	2048	1024
Ts	0.02844	0.2275	0.05689	0.1138	0.05689
Tg	2.81	22.45	5.6	11.22	5.6
Lamp (Up) (μs)	1	1	1	1	1
CS	1	1	1	1	1
CP	0.81	20.45	3.6	9.22	3.6
Ts+Tg	0.0313	0.25	0.0625	0.125	0.0625
Bit rate	4 Mbps	8-50 Mbps	4-25 Mbps	8-50 Mbps	8-50 Mbps

[0063] Here, Ts denotes a period of time indicating the length of a transmission symbol, Tg denotes a guard time, the unit of the CS is in μs, and bps indicates bits per second.

[0064] As is seen from Table 1, in an OFDM communication method and apparatus according to the present invention, the length of a transmission symbol Ts is adjusted by changing the number of carrier waves so that the method and apparatus can be adapted to various communication environments. The transmission symbol is adjusted to adapt to various communication environments for the following reasons.

[0065] For example, let's assume that a Veh B channel, from Tr 101 146 v3.0, which is disclosed in a book entitled "Digital Communications", written by J. Proakis, and published by McGraw Hill in 1995, is used; the number of carrier waves is 4096; a whole signal bandwidth is 18 MHz; and a spread factor (SF) is 4.

[0066] FIG. 16 is a graph showing changes in a bit error rate (BER) with respect to changes in Doppler frequency. The vertical axis indicates a BER, and the horizontal axis indicates Eb/No where Eb is energy per bit and No is the variance of noise.

[0067] The BER at a Doppler frequency, i.e., a channel change speed, of 170 (■) is greater than the BER at a channel change speed of 17 (★). The BER at a channel change speed of 500 (▲) is greater than the BER at the channel change speed of 170 (■). Consequently, as is seen from FIG. 16, the BER increases with an increase in a channel change speed.

[0068] In the meantime, when the same assumption as described above is adopted, with the exception that the SF is 1, the Doppler frequency is 500, and a Veh A channel is used instead of the Veh B channel, changes in a BER with respect to changes in the number of carrier waves will be described below. Here, the channels (Veh A and Veh B) are disclosed in Table 1.2.2.3 in page 43 of a book entitled "Selection Procedures for the Choice of Radio Transmission Technologies" and published by Universal Mobile Telecommunication System (UMTS), which is under a standardization group of European Telecommunications Standardization Institute (ETSI), in Technical Report (TR) 101112 of the ETSI.

[0069] FIG. 17 is a graph showing changes in a BER with respect to changes in the number of carrier waves. The vertical axis indicates a BER, and the horizontal axis indicates Eb/No.

[0070] As shown in FIG. 17, the BER when the number of carrier waves is 2048 (■) is greater than the BER when the number of carrier waves is 1024 (▲), and the BER when the number of carrier waves is 4096 (★) is greater than the BER when the number of carrier waves is 2048 (■). Consequently, as is seen from FIG. 17, when the length of a transmission symbol is decreased by decreasing the number of carrier waves from 4096 to 1096, the influence of the Doppler frequency is reduced, thereby decreasing the BER. Accordingly, if transmission data is repeated two times, as shown in FIG. 10 or 11, influence due to a change in a channel length is decreased and interchannel interference is prevented.

[0071] As described above, in an OFDM communication method and apparatus to adapt to channel characteristics according to the present invention, at least one of the length and the format of a transmission symbol and the format of a frame is changed to adapt to channel characteristics such as a channel change speed and a channel length so that communication can be accomplished at a low BER and high efficiency under various environments and a terminal can be simply implemented. In particular, under an environment in which a channel change speed is fast and a channel length is long, communication reliability can be enhanced. Since the transmission symbol includes the first symbol regardless of a cell radius, as shown in FIGS. 4 through 6, the present invention facilitates wireless resource management such as association and handover.

[0072] While the present invention has been particularly shown and described with reference to exemplary embodiments thereof, it will be understood by those of ordinary skill in the art that various changes in form and details may be therein without departing from the scope of the invention as defined by the appended claims.

Claims

1. An orthogonal frequency division multiplexing (OFDM) communication method to adapt to channel characteristics, comprising the steps of changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol depending on a type of the transmission symbol and a radius of a cell, in which communication is performed.

2. The OFDM communication method of claim 1, wherein the changing steps comprises the following steps:

   (a) determining whether the transmission symbol is a symbol that is used for a control channel;
   (b) if it is determined that the transmission symbol is the symbol that is used for a control channel, determining a first symbol containing control information as the transmission symbol;
   (c) if it is determined that the transmission symbol is not the symbol that is used for a control channel, determining whether the cell radius is greater than a first predetermined value;
   (d) if it is determined that the cell radius is greater than the first predetermined value, determining a second symbol, which is suitable to channel characteristics where a channel change speed is slow and a channel length is long, or a third symbol, which is suitable to channel characteristics where the channel change speed is fast and the channel length is long, as the transmission symbol;
   (e) if it is determined that the cell radius is not greater than the first predetermined value, determining whether the cell radius is greater than a second predetermined value;
   (f) if it is determined that the cell radius is greater than the second predetermined value, determining a fourth symbol, which is suitable to channel characteristics where the channel change speed and the channel length are medium, as the transmission symbol; and
   (g) if it is determined that the cell radius is not greater than the second predetermined value, determining a fifth symbol, which is suitable to channel characteristics where the channel change speed is slow and the channel length is short, as the transmission symbol,

   wherein the second predetermined value is less than the first predetermined value, a length of the fourth symbol is less than a length of the second symbol, and a length of each of the first, third, and fifth symbols is less than the length of the fourth symbol.

3. The OFDM communication method of claim 2, wherein the length of each of the second, third, fourth, and fifth symbols is an integer multiple of the length of the first symbol.

4. The OFDM communication method of claim 2, wherein the length of each of the second, third, and fourth symbols is an integer multiple of the length of the fifth symbol.

5. The OFDM communication method of any one of claims 2 to 4, further comprising the step of adjusting the length of the determined transmission symbol by changing the number of carrier waves

6. The OFDM communication method of claim 2, wherein step (d) comprises determining the second or third symbol as the transmission symbol and converting the format of the frame into a macro format if it is determined that the cell radius is greater than the first predetermined value,

step (f) comprises determining the fourth symbol as the transmission symbol and converting the format of the frame into a micro format if it is determined that the cell radius is greater than the second predetermined value, and

step (g) comprises determining the fifth symbol as the transmission symbol and converting the format of the frame into a pico format if it is determined that the cell radius is not greater than the second predetermined value.

7. The OFDM communication method of claim 1, wherein changing step comprises the steps of:

(h) determining whether the radius cell is greater than a first predetermined value;
(i) if it is determined that the radius cell is greater than the first predetermined value, converting the format of the frame into a macro format;
(j) if it is determined that the radius cell is not greater than the first predetermined value, determining whether the radius cell is greater than a second predetermined value;
(k) if it is determined that the radius cell is greater than the second predetermined value, converting the format of the frame into a micro format; and
(l) if it is determined that the radius cell is not greater than the second predetermined value, converting the format of the frame into a pico format,

wherein the first predetermined value is greater than the second predetermined value.

8. The OFDM communication method of claim 7, wherein the macro format comprises: a first symbol, which contains control information;

a second symbol, which is suitable to channel characteristics where a channel change speed is slow and a channel length is long; and

a third symbol, which is suitable to channel characteristics where the channel change speed is fast and the channel length is long.

9. The OFDM communication method of claim 7 or 8, wherein the micro format comprises:

a first symbol, which contains control information; and
a fourth symbol, which is suitable to channel characteristics where a channel change speed and a channel length are medium.

10. The OFDM communication method of claim 7, 8 or 9, wherein the pico format comprises:

a first symbol, which contains control information; and
a fifth symbol, which is suitable to channel characteristics where a channel change speed is slow and a channel length is short.

11. The OFDM communication method of claim 2, wherein step (d) further comprises the steps of:

(d1) if it is determined that the cell radius is greater than the first predetermined value, determining whether the channel change speed is greater than a predetermined speed;
(d2) if it is determined that the channel change speed is not greater than the predetermined speed, determining the second symbol as the transmission symbol; and
(d3) if it is determined that the channel change speed is greater than the predetermined speed, determining the third symbol as the transmission symbol.

12. The OFDM communication method of claim 11, wherein the second symbol determined as the transmission symbol in step (d2) comprises:

a first cyclic prefix, which contains an end portion of transmission data;

a first transmission signal, which contains the transmission data; and
a first cyclic suffix, which contains a beginning portion of the transmission data.

13. The OFDM communication method of claim 11, wherein the third symbol determined as the transmission symbol in step (d3) comprises:

a first cyclic prefix, which contains a plurality of end portions of transmission data and a beginning portion of the transmission data;
a first transmission signal, which contains the transmission data;
a second transmission signal, which contains the transmission data; and
a first cyclic suffix, which contains the beginning portion of the transmission data.

14. The OFDM communication method of claim 11, wherein the third symbol comprises:

a first cyclic prefix, which contains a plurality of end portions of transmission data;
a first transmission signal, which contains the transmission data;
a second transmission signal, which contains the transmission data; and
a first cyclic suffix, which contains a plurality of beginning portions of the transmission data.

15. An orthogonal frequency division multiplexing (OFDM) communication apparatus to adapt to channel characteristics, comprising:

a symbol inspector, for inspecting a type of a transmission symbol and outputting the result of the inspection as a first control signal; and
a symbol and format converter, for changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol in response to the first control signal and a radius of a cell, in which communication is performed.

16. The OFDM communication apparatus of claim 15, wherein the symbol and format converter comprises:

a first comparator, for comparing the cell radius with a first predetermined value in response to the first control signal and outputting the result of the comparison as a second control signal;
a second comparator, for comparing the cell radius with a second predetermined value in response to the second control signal and outputting the result of the comparison as a third control signal; and
a first converter, for determining one among first, second, third, fourth, and fifth symbols as the transmission symbol in response to the first, second, and third control signals and outputting the determined symbol,

wherein the second predetermined value is less than the first predetermined value, the first symbol contains control information, the second symbol is suitable to channel characteristics where a channel change speed is slow and a channel length is long, the third symbol is suitable to channel characteristics where the channel change speed is fast and the channel length is long, the fourth symbol is suitable to channel characteristics where the channel change speed and the channel length are medium, and the fifth symbol is suitable to channel characteristics where the channel change speed is slow and the channel length is short.

17. The OFDM communication apparatus of claim 15 or 16, wherein the symbol and format converter comprises:

a third comparator, for comparing the cell radius with a first predetermined value and outputting the result of the comparison as a fourth control signal;
a fourth comparator, for comparing the cell radius with a second predetermined value in response to the fourth control signal and outputting the result of the comparison as a fifth control signal; and
a second converter, for converting the format of the frame into one of a macro format, a micro format, and a pico format in response to the fourth and fifth control signals,

wherein the first predetermined value is greater than the second predetermined value.

18. The OFDM communication apparatus of claim 16, wherein the first converter converts the format of the frame into one of a macro format, a micro format, and a pico format in response to the second and third control signals.

19. The OFDM communication apparatus of claim 16, wherein the first converter comprises a fifth comparator, for comparing the channel change speed with a predetermined speed in response to the second control signal and outputting the result of the comparison as a sixth control signal, and a format converter, for converting the format of the determined symbol in response to the sixth control signal.

FIG. 1

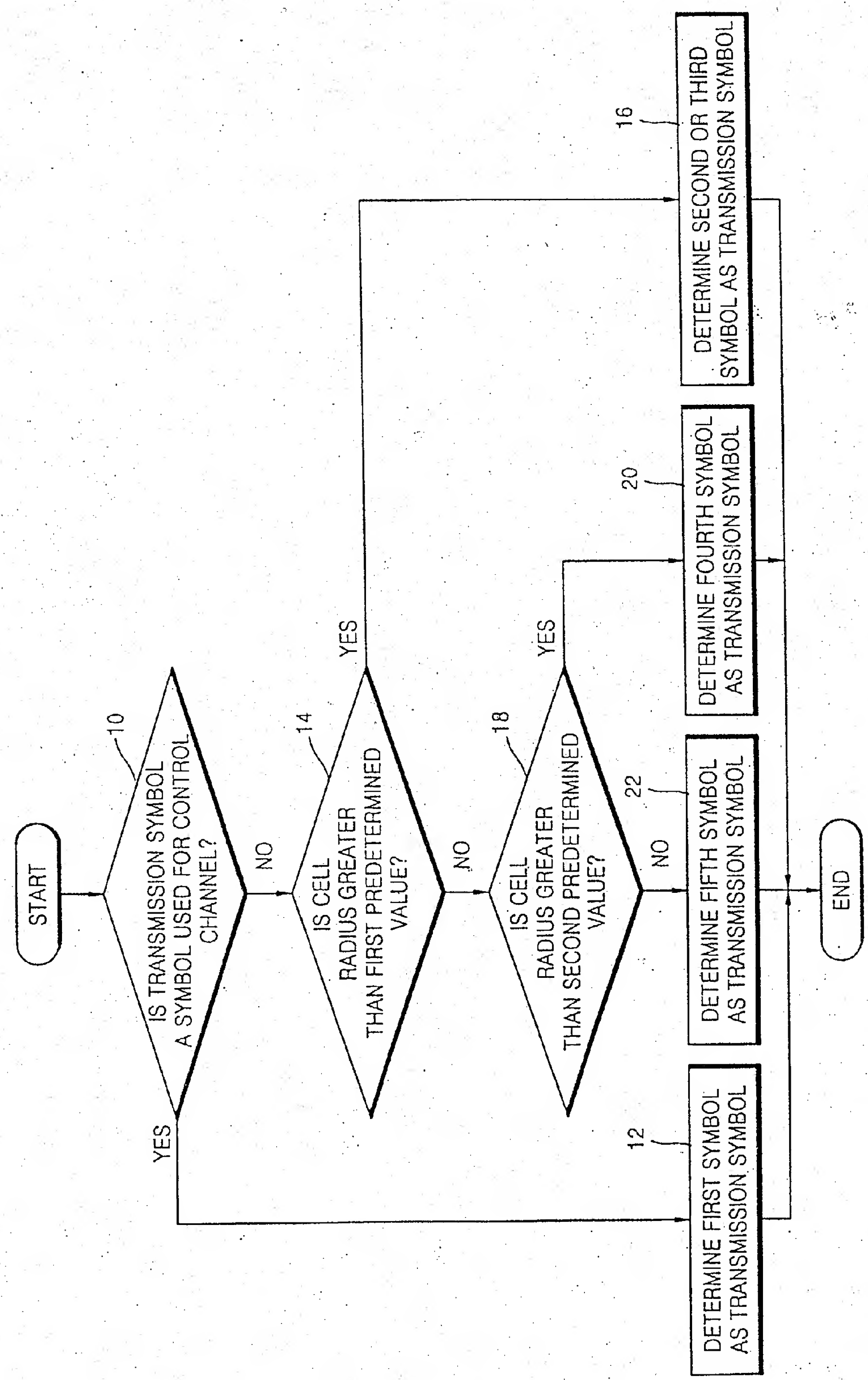

START
IS TRANSMISSION SYMBOL A SYMBOL USED FOR CONTROL CHANNEL?
10
YES
NO
DETERMINE FIRST SYMBOL AS TRANSMISSION SYMBOL
12
IS CELL RADIUS GREATER THAN FIRST PREDETERMINED VALUE?
14
YES
NO
DETERMINE SECOND OR THIRD SYMBOL AS TRANSMISSION SYMBOL
16
IS CELL RADIUS GREATER THAN SECOND PREDETERMINED VALUE?
18
YES
NO
DETERMINE FOURTH SYMBOL AS TRANSMISSION SYMBOL
20
DETERMINE FIFTH SYMBOL AS TRANSMISSION SYMBOL
22
END

FIG. 2

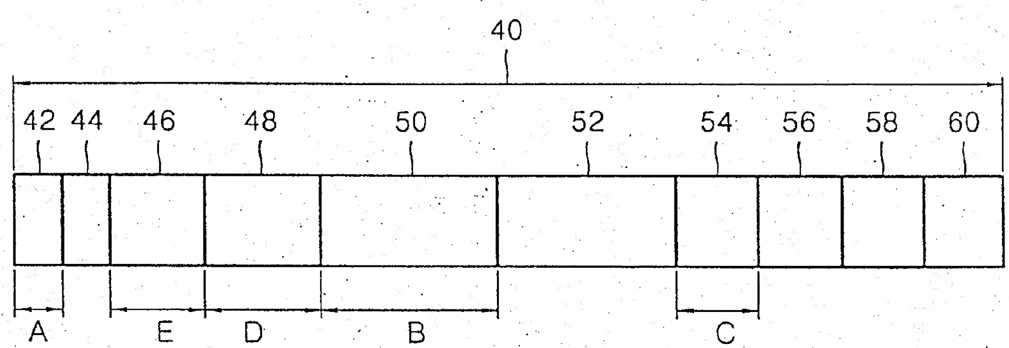


FIG. 3

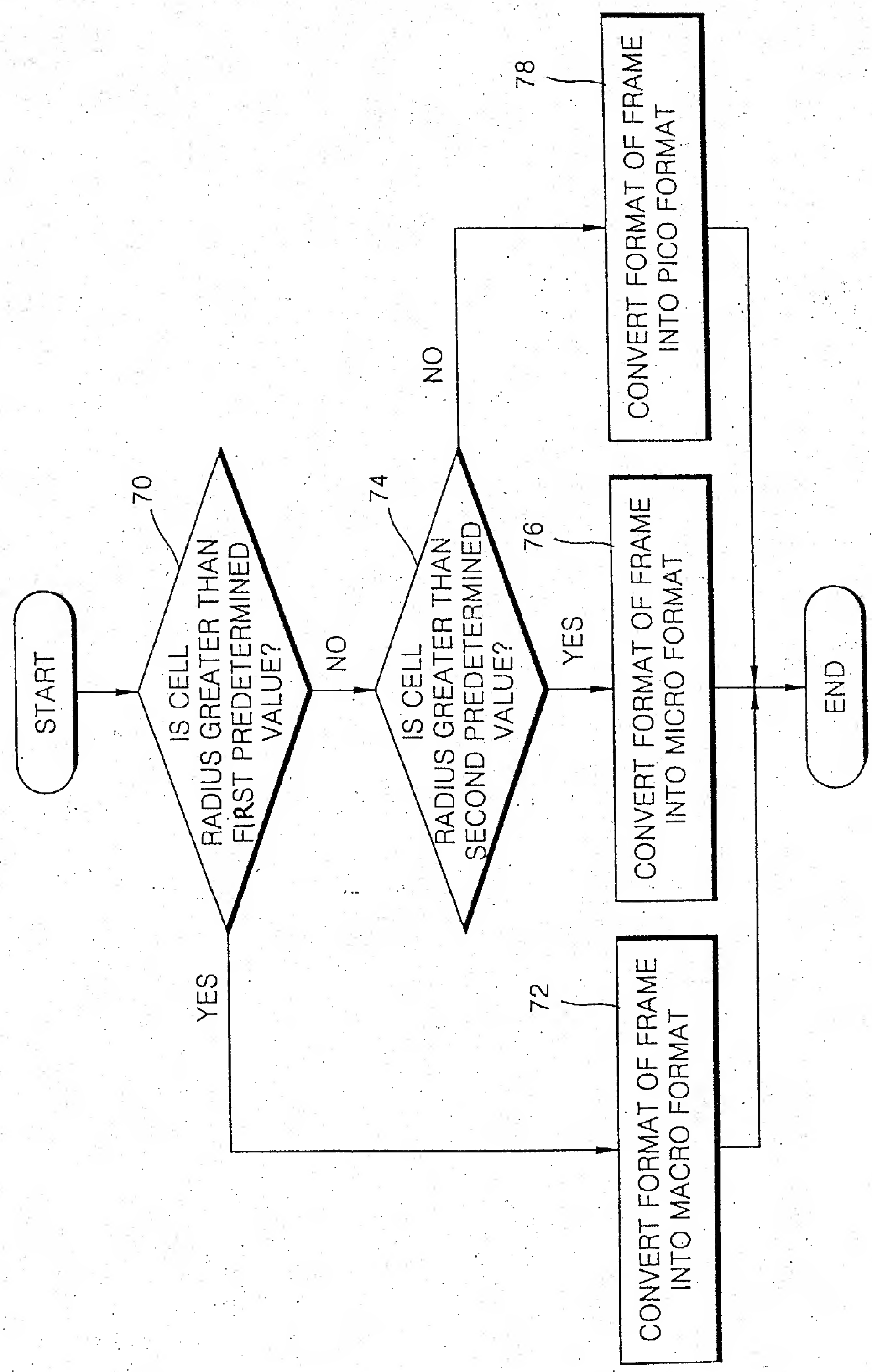

START
IS CELL RADIUS GREATER THAN FIRST PREDETERMINED VALUE?
70
YES
NO
IS CELL RADIUS GREATER THAN SECOND PREDETERMINED VALUE?
74
YES
NO
CONVERT FORMAT OF FRAME INTO MACRO FORMAT
72
CONVERT FORMAT OF FRAME INTO MICRO FORMAT
76
CONVERT FORMAT OF FRAME INTO PICO FORMAT
78
END

FIG. 4

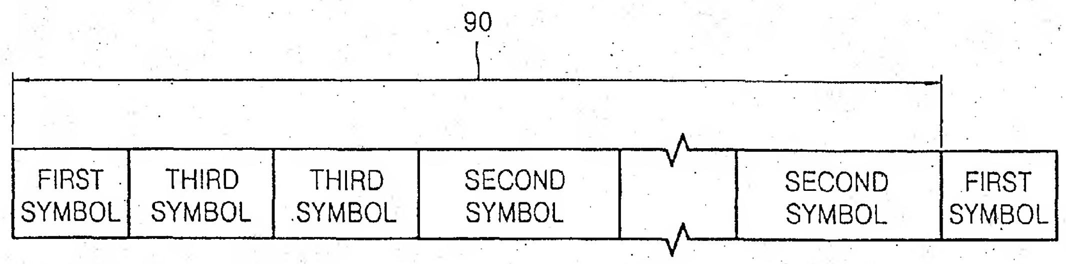


FIG. 5

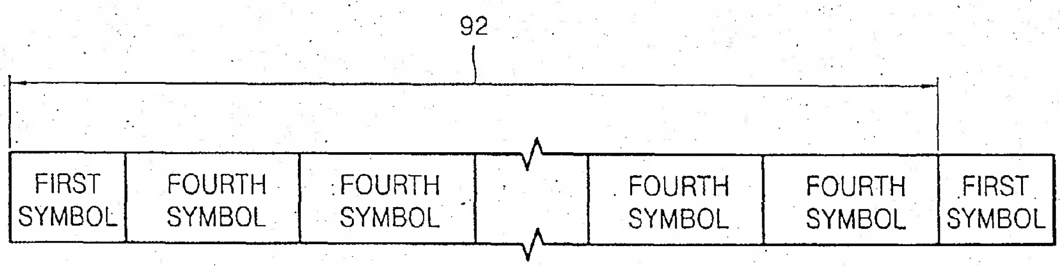


FIG. 6

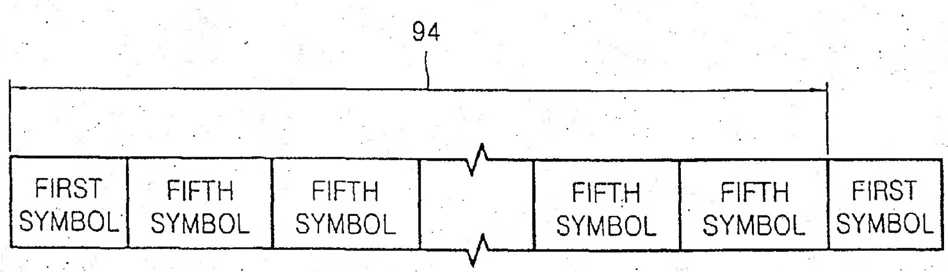


FIG. 7

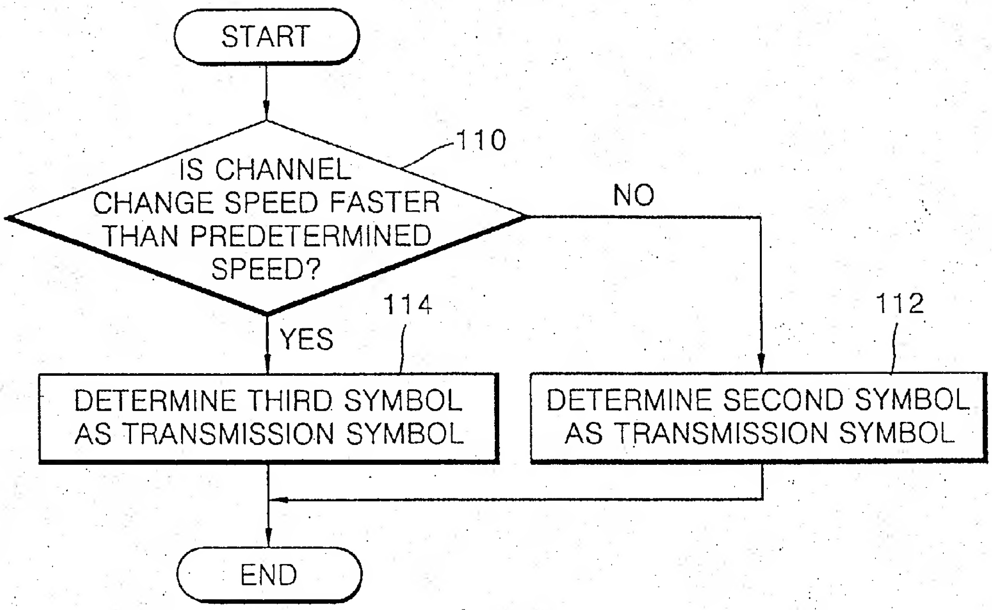

START
IS CHANNEL CHANGE SPEED FASTER THAN PREDETERMINED SPEED?
110
NO
YES
114
112
DETERMINE THIRD SYMBOL AS TRANSMISSION SYMBOL
DETERMINE SECOND SYMBOL AS TRANSMISSION SYMBOL
END

FIG. 8

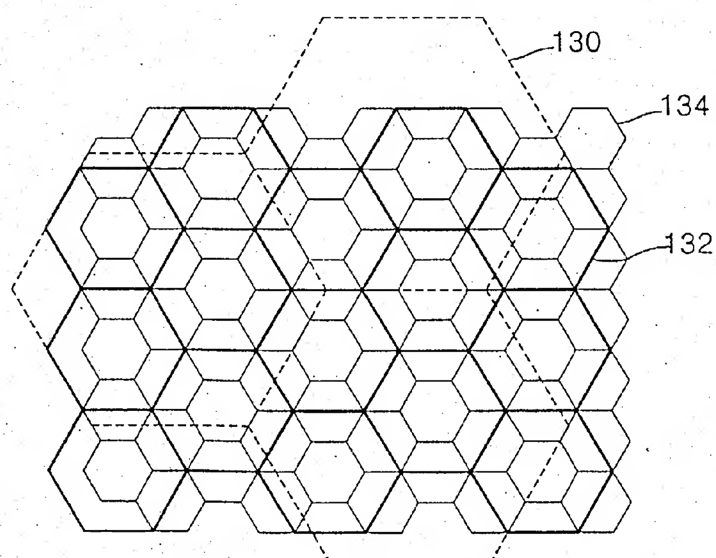

130
134
132

FIG. 9

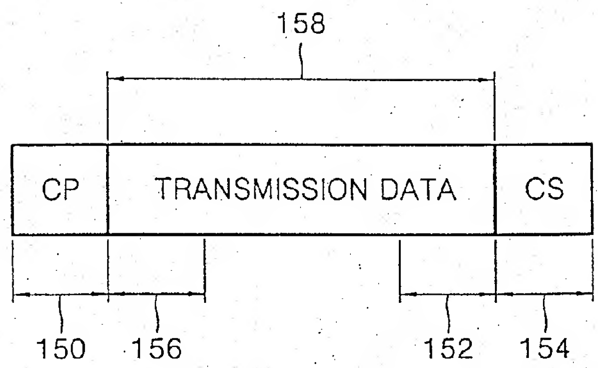


FIG. 10

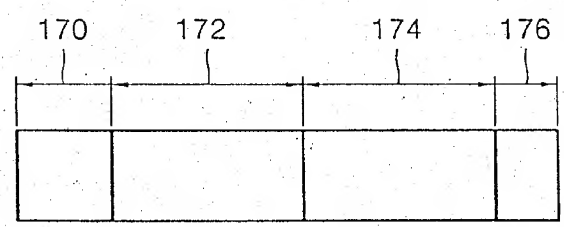


FIG. 11

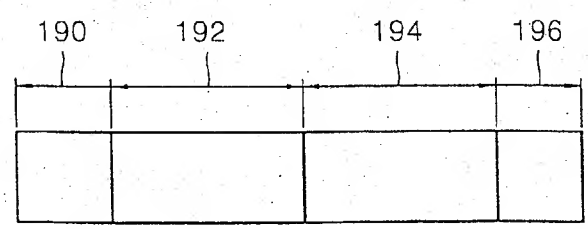


FIG. 12

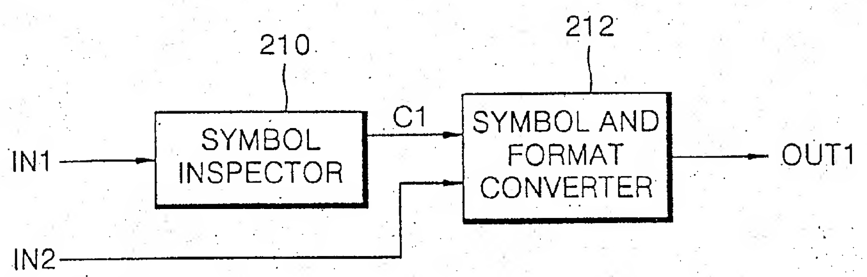


FIG. 13

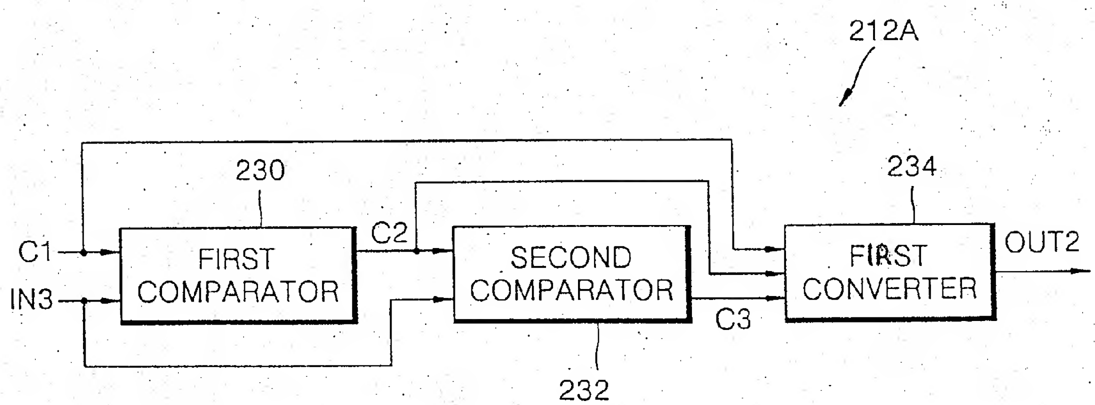


FIG. 14

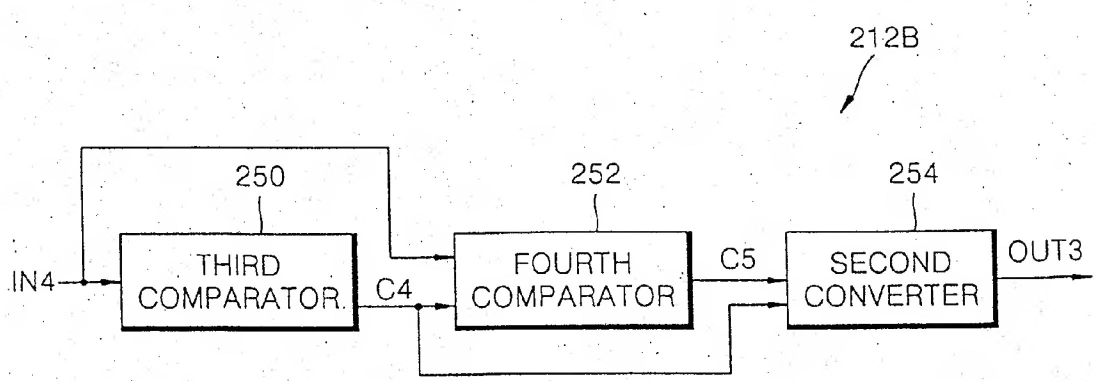


FIG. 15

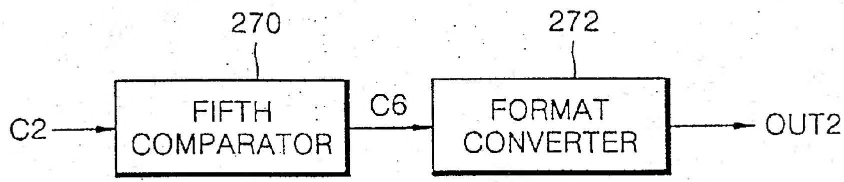


FIG. 16

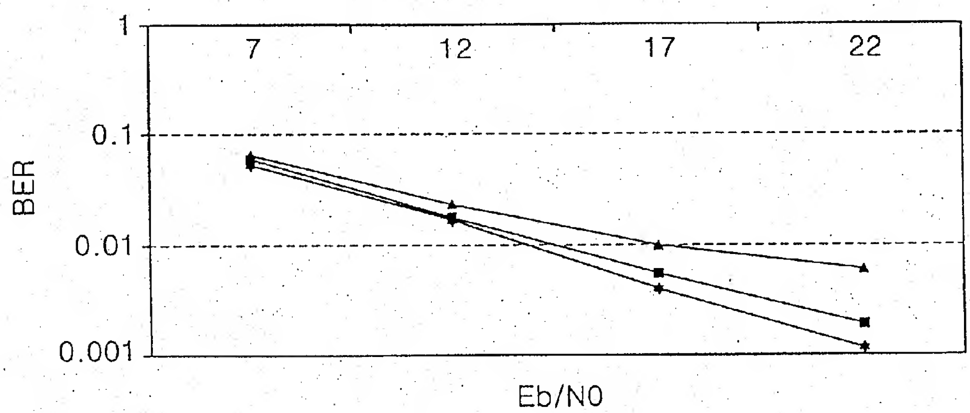
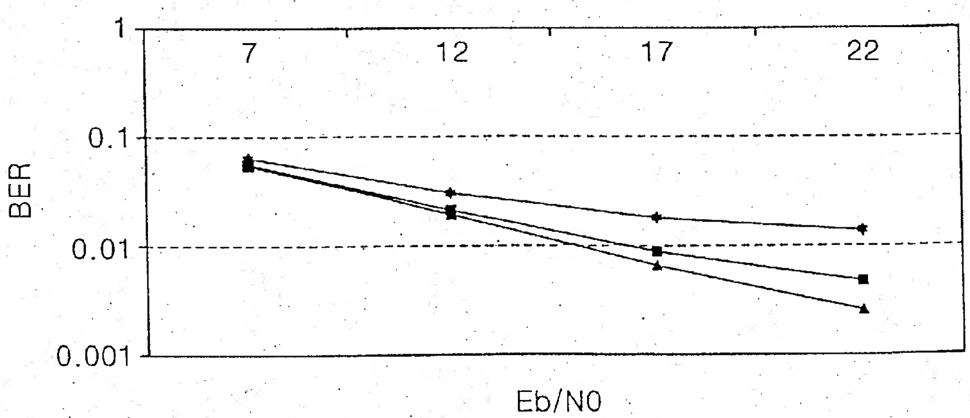


FIG. 17



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DATA DETECTION AND DEMODULATION FOR WIRELESS COMMUNICATION SYSTEMS

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Inventor(s): WALTON J RODNEY [US]; WALLACE MARK S [US] +

Applicant(s): QUALCOMM INC [US] +

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- WO2004038989 (A3)
- MXPA05004311 (A)
- KR20050071576 (A)
- JP2010213309 (A)

more >>

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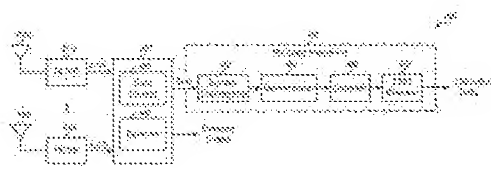
- EP0772329 (A1)
- EP0895387 (A1)
- US6314113 (B1)
- US5959965 (A)
- JP11205273 (A)

Abstract not available for EP 1556984 (A2)
Abstract of corresponding document: **WO 2004038989 (A2)**

Techniques for detecting and demodulating data transmissions in wireless communication systems. In one aspect, a decision-directed detector detects for data transmissions in a received signal by utilizing received data symbols as well as received pilot symbols. The decision-directed detector may be designed to perform differential detection in the frequency domain or coherent detection in the time domain, and may be used with multi-carrier modulation (e.g., OFDM). In another aspect, an adaptive threshold is used to perform detection of received data transmissions. A threshold may be determined for each data transmission hypothesized to have been received. The threshold may be computed, for example, based on the signal plus noise energy of the hypothesized data transmission.



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60/432,626	10 December 2002 (10.12.2002)	US
10/692,833	23 October 2003 (23.10.2003)	US

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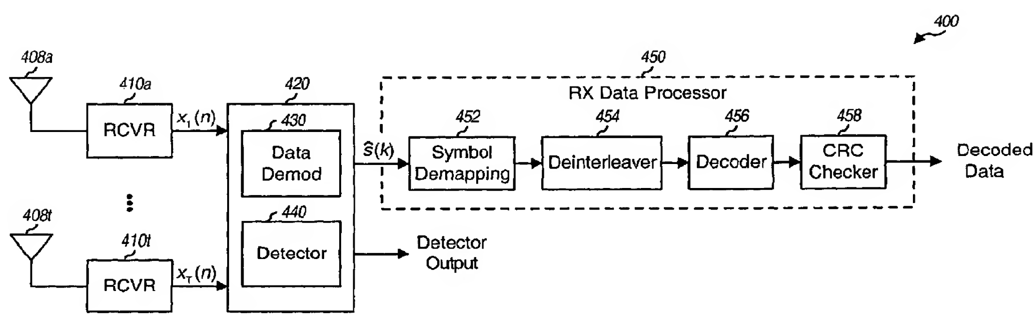

(57) **Abstract:** Techniques for detecting and demodulating data transmissions in wireless communication systems. In one aspect, a decision-directed detector detects for data transmissions in a received signal by utilizing received data symbols as well as received pilot symbols. The decision-directed detector may be designed to perform differential detection in the frequency domain or coherent detection in the time domain, and may be used with multi-carrier modulation (e.g., OFDM). In another aspect, an adaptive threshold is used to perform detection of received data transmissions. A threshold may be determined for each data transmission hypothesized to have been received. The threshold may be computed, for example, based on the signal plus noise energy of the hypothesized data transmission.

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DATA DETECTION AND DEMODULATION FOR WIRELESS COMMUNICATION SYSTEMS

Claim of Priority under 35 U.S.C. §119

[0001] This application claims the benefit of U.S. Provisional Application Serial No. 60/421,309, entitled "MIMO WLAN System," filed on October 25, 2002, assigned to the assignee of the present application, and incorporated herein by reference in its entirety for all purposes.

[0002] This application claims the benefit of U.S. Provisional Application Serial No. 60/432,626, entitled "Data Detection and Demodulation for Wireless Communication Systems," filed on December 10, 2002, assigned to the assignee of the present application, and incorporated herein by reference in its entirety for all purposes.

BACKGROUND

I. Field

[0003] The present invention relates generally to data communication, and more specifically to techniques for detecting and demodulating data transmissions in wireless communication systems.

II. Background

[0004] In a wireless communication system, data to be transmitted is typically processed (e.g., coded and modulated) and then upconverted onto a radio frequency (RF) carrier signal to generate an RF modulated signal that is more suitable for transmission over a wireless channel. The RF modulated signal is then transmitted from a transmitter and may reach a receiver via a number of propagation paths in the wireless channel. The characteristics of the propagation paths typically vary over time due to a number of factors such as, for example, fading, multipath, and external interference. Consequently, the RF modulated signal may experience different channel conditions (e.g., different fading and multipath effects) and may be associated with different complex gains across the operating bandwidth of the system.

[0005] To achieve high performance, a pilot (i.e., a reference signal) is often transmitted by the transmitter to assist the receiver in performing a number of functions. The pilot is typically generated based on known symbols and processed in a known manner. The pilot may be used by the receiver for channel estimation, timing and frequency acquisition, coherent demodulation, and so on

[0006] It is often desirable or necessary to detect for the presence of data transmissions in a received signal. The detection for data transmissions is normally achieved by processing the pilot for each data transmission hypothesized to have been received. If the energy of the pilot is greater than a particular threshold, then the hypothesized data transmission is further processed (e.g., demodulated and decoded). An error detection code, such as a cyclic redundancy check (CRC), is then typically relied upon to determine whether the data transmission was decoded correctly or in error.

[0007] In some wireless communication systems, detection based on the pilot alone is not sufficient. This may be the case, for example, when operating at a low received signal-to-noise ratio (SNR). Moreover, an error detection code may not be available for use to verify the correctness of the received data transmission.

[0008] There is therefore a need in the art for techniques to detect and demodulate data transmissions in such wireless communication systems.

SUMMARY

[0009] Techniques are provided herein for detecting and demodulating data transmissions in wireless communication systems. In one aspect, a decision-directed detector is provided to detect for data transmissions in a received signal. This detector utilizes received data symbols as well as received pilot symbols to perform the detection and is thus able to provide improved detection performance. The decision-directed detector may be designed to operate in the frequency domain or the time domain. For a system utilizing multi-carrier modulation (e.g., OFDM), the detector may be designed to perform differential detection in the frequency domain or coherent detection in the time domain, both of which are described in detail below.

[0010] In another aspect, an adaptive threshold is used to perform detection of received data transmissions. A threshold may be determined for each data transmission hypothesized to have been received. The threshold may be computed, for example,

based on the total received signal energy (i.e., signal plus noise plus interference) of the hypothesized data transmission. The use of an adaptive threshold can provide robust detection performance in many operating environments, such as in an unlicensed frequency band where various sources of interference may be present.

[0011] Various aspects and embodiments of the invention are described in further detail below. For example, receiver structures for various transmission schemes are also described herein.

BRIEF DESCRIPTION OF THE DRAWINGS

[0012] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[0013] FIG. 1 shows a wireless communication system;

[0014] FIGS. 2A and 2B show exemplary protocol data units (PDUs) for Channels 1 and 2, respectively;

[0015] FIG. 3A shows a block diagram of a transmitter unit;

[0016] FIG. 3B illustrates an OFDM symbol;

[0017] FIG. 4 shows a block diagram of a receiver unit;

[0018] FIG. 5 shows a correlation detector;

[0019] FIG. 6 shows an embodiment of the correlation detector;

[0020] FIG. 7 shows a detector/data demodulator that includes a data demodulator and a decision-directed detector;

[0021] FIG. 8A shows an embodiment of the data demodulator;

[0022] FIG. 8B shows a decision-directed detector that performs differential detection in the frequency domain;

[0023] FIG. 8C shows a decision-directed detector that performs coherent detection in the time domain;

[0024] FIG. 9 shows a block diagram of an access point and a user terminal;

[0025] FIGS. 10A and 10B show exemplary transmissions on Channels 1 and 2, respectively; and

[0026] FIGS. 11A and 11B show the receiver processing for Channels 1 and 2, respectively.

DETAILED DESCRIPTION

[0027] The word "exemplary" is used herein to mean "serving as an example, instance, or illustration." Any embodiment or design described herein as "exemplary" is not necessarily to be construed as preferred or advantageous over other embodiments or designs.

[0028] FIG. 1 shows a wireless communication system 100 that includes a number of access points (APs) 110 that communicate with a number of user terminals (UTs) 120. (For simplicity, only one access point is shown in FIG. 1.) An access point may also be referred to as a base station or some other terminology. Each user terminal may be a fixed or mobile terminal and may also be referred to as an access terminal, a mobile station, a remote station, a user equipment (UE), a wireless device, or some other terminology. Each user terminal may communicate with one or possibly multiple access points on the downlink and/or the uplink at any given moment. The downlink (i.e., forward link) refers to transmission from the access point to the user terminal, and the uplink (i.e., reverse link) refers to transmission from the user terminal to the access point.

[0029] The techniques described herein for detecting and demodulating data transmission may be used for various wireless communication systems. For example, these techniques may be used for systems that employ (1) one or multiple antennas for data transmission and one or multiple antennas for data reception, (2) various modulation techniques (e.g., CDMA, OFDM, and so on), and (3) one or multiple frequency bands for the downlink and uplink.

[0030] For clarity, the techniques are specifically described below for an exemplary wireless communication system. In this system, a receiver is equipped with multiple (T) antennas for data reception, and a transmitter may be equipped with one or multiple antennas. The system further employs orthogonal frequency division multiplexing (OFDM), which effectively partitions the overall system bandwidth into multiple (N) orthogonal subbands. For OFDM, the data or pilot to be transmitted on each subband is first modulated (i.e., symbol mapped) using a particular modulation scheme. Signal

values of zero are provided for subbands not used for data/pilot transmission. For each OFDM symbol period, the modulation symbols and zero signal values for all N subbands are transformed to the time domain using an inverse fast Fourier transform (IFFT) to obtain a transformed symbol that comprises N time-domain samples. To combat inter-symbol interference (ISI), a portion of each transformed symbol is often repeated to form a corresponding OFDM symbol, which is then transmitted over the wireless channel. An OFDM symbol period (or simply, a symbol period) corresponds to the duration of one OFDM symbol, which is the smallest unit of transmission for the system. In one specific design, the system bandwidth is 20 MHz, $N = 64$, the subbands are assigned indices of -32 to +31, the duration of each transformed symbol is 3.2 μsec, the cyclic prefix is 800 nsec, and the duration of each OFDM symbol is 4.0 μsec.

[0031] For clarity, two specific transmission schemes and two receiver structures are described below. The first transmission scheme is used for Transport Channel 1 (or simply, Channel 1 or CH1) and has the following characteristics: (1) transmissions on Channel 1 are not time-compensated at the transmitter and arrive at unknown times at the receiver, and (2) each transmission on Channel 1 includes multiple OFDM symbols for data and pilot. The second transmission scheme is used for Transport Channel 2 (or simply, Channel 2 or CH2) and has the following characteristics: (1) transmissions on Channel 2 are time-compensated at the transmitter and arrive time-aligned to slot boundaries at the receiver, and (2) each transmission on Channel 2 includes a single OFDM symbol for both data and pilot. Slow and fast random access channels with similar characteristics as those of Channels 1 and 2 are described in the aforementioned U.S. Patent Application Serial No. 60/432,440.

[0032] FIG. 2A shows an exemplary protocol data unit (PDU) 210 that may be used for Channel 1 (CH1 PDU). CH1 PDU 210 comprises a reference portion 220 that is time division multiplexed (TDM) with a CH1 message portion 230. Reference portion 220 includes P pilot OFDM symbols 222, where P can be any integer one or greater. The pilot OFDM symbols are used to facilitate acquisition and detection of a CH1 transmission as well as to aid in coherent demodulation of the CH1 message portion. CH1 message portion 230 includes D data OFDM symbols 232, where D can be any integer one or greater. The pilot and data OFDM symbols may be generated as described below.

[0033] FIG. 2B shows an exemplary PDU 250 that may be used for Channel 2 (CH2 PDU). CH2 PDU 250 comprises a reference portion 260 that is subband multiplexed with a CH2 message portion 270. Reference portion 260 comprises a set of pilot symbols that is transmitted on one set of subbands (shown as shaded subbands in FIG. 2B). CH2 message portion 270 comprises a group of data symbols that is transmitted on another set of subbands. The data symbols are generated by coding, interleaving, and symbol mapping a CH2 message 280. The frequency-domain multiplexed pilot and data symbols are processed to generate time-domain CH2 PDU 250, as described below.

[0034] In the embodiment shown in FIG. 2B, the pilot subbands and data subbands are interlaced such that each data subband is flanked on both sides by pilot subbands. The pilot symbols transmitted on the pilot subbands may be used to estimate the channel responses for the data subbands and for coherent demodulation. Other subband multiplexing schemes may also be implemented, and this is within the scope of the invention. For example, each group of Q data subbands may be flanked on both sides by pilot subbands, where Q may be any positive integer.

[0035] FIG. 3A shows a block diagram of an embodiment of a transmitter unit 300 that can perform transmit data processing for Channels 1 and 2 described above. Transmitter unit 300, which may be implemented within an access point or a user terminal, includes a transmit (TX) data processor 310, an optional TX spatial processor 330, and one OFDM modulator 340 for each transmit antenna.

[0036] Within TX data processor 310, a CRC generator 312 receives data for a CH1 or CH2 message and (optionally) generates a CRC value for the message. An encoder 314 then codes the message data and the CRC value (if included) in accordance with a particular coding scheme to provide code bits. An interleaver 316 next interleaves (i.e., reorders) the code bits based on a particular interleaving scheme to provide frequency and possibly time diversity. A symbol mapping unit 318 then maps the interleaved data in accordance with a particular modulation scheme to provide modulation symbols, which are also referred to as data symbols and denoted as $s(k)$.

[0037] A multiplexer (MUX) 320 receives and multiplexes the data symbols with pilot symbols in the manner defined for the CH1 or CH2 message being processed. For the embodiment shown in FIG. 2A, a CH1 PDU comprises P pilot OFDM symbols followed by D data OFDM symbols. For a CH1 message, multiplexer 320 provides a

set of pilot symbols $\{p_1(k)\}$ for each of the P pilot OFDM symbols, then the data symbols for each of the D data OFDM symbols. For the embodiment shown in FIG. 2B, a CH2 PDU comprises L+1 pilot symbols interlaced with L data symbols. For a CH2 message, multiplexer 320 provides a set of L+1 pilot symbols $\{p_2(k)\}$ multiplexed with a group of L data symbols. In any case, multiplexer 320 provides a stream of multiplexed data and pilot symbols.

[0038] Table 1 shows a specific embodiment of two sets of pilot symbols, $\{p_1(k)\}$ and $\{p_2(k)\}$, for CH1 and CH2 reference portions. In this embodiment, only 52 of the 64 total subbands are used for data and pilot transmission, and the other 12 subbands (with zero entries in Table 1) are not used. In an embodiment, the pilot symbols are QPSK modulation symbols. The 52 pilot symbols for the CH1 reference portion are selected such that a waveform generated based on these pilot symbols has minimum peak-to-average variation. This characteristic allows the pilot OFDM symbol to be transmitted at a higher power level, which can provide improved performance.

Table 1 - Pilot Symbols for CH1 and CH2

Sub-band Index	**CH1 Pilot Symbol** $p_1(k)$	**CH2 Pilot Symbol** $p_2(k)$	**Sub-band Index**	**CH1 Pilot Symbol** $p_1(k)$	**CH2 Pilot Symbol** $p_2(k)$	**Sub-band Index**	**CH1 Pilot Symbol** $p_1(k)$	**CH2 Pilot Symbol** $p_2(k)$	**Sub-band Index**	**CH1 Pilot Symbol** $p_1(k)$	**CH2 Pilot Symbol** $p_2(k)$
-32	0	0	-16	$-1+j$	data	0	0	0	16	$-1+j$	data
-31	0	0	-15	$1-j$	$1+j$	1	$1-j$	$-1-j$	17	$-1+j$	$1-j$
-30	0	0	-14	$1+j$	data	2	$-1-j$	data	18	$1-j$	data
-29	0	0	-13	$1-j$	$1+j$	3	$-1-j$	$-1-j$	19	$1+j$	$-1-j$
-28	0	0	-12	$1-j$	data	4	$-1-j$	data	20	$-1+j$	data
-27	0	0	-11	$-1-j$	$1+j$	5	$-1+j$	$1+j$	21	$1+j$	$-1-j$
-26	$-1-j$	$-1+j$	-10	$-1-j$	data	6	$1+j$	data	22	$-1+j$	data
-25	$-1+j$	$-1+j$	-9	$1-j$	$1-j$	7	$-1-j$	$-1-j$	23	$1+j$	$-1-j$
-24	$-1+j$	data	-8	$-1-j$	data	8	$-1+j$	data	24	$-1+j$	data
-23	$-1+j$	$-1-j$	-7	$1+j$	$-1+j$	9	$-1-j$	$1-j$	25	$1-j$	$-1+j$
-22	$1-j$	data	-6	$-1+j$	data	10	$-1-j$	data	26	$-1-j$	$1-j$
-21	$1-j$	$-1-j$	-5	$-1-j$	$-1-j$	11	$1+j$	$1+j$	27	0	0

-20	1+j	data	-4	-1+j	data	12	1-j	data	28	0	0
-19	-1-j	-1-j	-3	-1+j	-1+j	13	-1+j	1-j	29	0	0
-18	-1+j	data	-2	1-j	data	14	-1-j	data	30	0	0
-17	1+j	1+j	-1	-1+j	-1+j	15	1+j	-1+j	31	0	0

[0039] If multiple antennas are available, then an optional TX spatial processor 330 may be used to perform spatial processing on the multiplexed data and pilot symbols. For example, TX spatial processor 330 may perform spatial processing for (1) beam-steering or beam-forming to transmit the symbols on a single spatial channel of a MIMO channel, (2) transmit diversity to transmit the symbols on multiple antennas and subbands to achieve diversity, or (3) spatial multiplexing to transmit the symbols on multiple spatial channels. Spatial processing for all of these transmission modes is described in detail in the aforementioned provisional U.S. Application Serial No. 60/421,309.

[0040] TX spatial processor 330 provides one stream of transmit symbols for each antenna. The transmit symbols are simply the multiplexed data and pilot symbols if spatial processing is not performed. Each transmit symbol stream is provided to a respective OFDM modulator 340. Within each OFDM modulator 340, an inverse fast Fourier transform (IFFT) unit 342 converts each sequence of N transmit symbols into a time-domain transformed symbol comprised of N time-domain samples, where N is the total number of subbands. For each transformed symbol, a cyclic prefix generator 344 repeats a portion of the transformed symbol to form a corresponding OFDM symbol comprised of M samples. Cyclic prefix generator 344 provides a stream of OFDM symbols to a transmitter (TMTR) 346, which converts the OFDM symbol stream into one or more analog signals and further amplifies, filters, and frequency upconverts the analog signal(s) to generate an RF modulated signal that is then transmitted from an associated antenna 350.

[0041] **FIG. 3B** illustrates an OFDM symbol, which is composed of two parts: a cyclic prefix and a transformed symbol. In an embodiment, $N = 64$, the cyclic prefix comprises 16 samples, and each OFDM symbol comprises $M = 80$ samples. The cyclic prefix is a copy of the last 16 samples (i.e., a cyclic continuation) of the transformed symbol and is inserted in front of the transformed symbol. The cyclic prefix ensures

that the OFDM symbol retains its orthogonal property in the presence of multipath delay spread.

[0042] **FIG. 10A** shows an exemplary transmission on Channel 1. The time line for Channel 1 is divided into CH1 slots, with each CH1 slot having a particular duration (e.g., P+D OFDM symbol periods). In an embodiment, one CH1 PDU may be transmitted on each CH1 slot.

[0043] User terminals A and B have locked their timing and frequency to that of the system. This may be achieved by receiving a transmission (e.g., a beacon pilot) that carries or is embedded with timing information. The user terminals then set their timing based on the received timing information. However, the timing of each user terminal may be skewed (or delayed) with respect to the system timing, where the amount of skew typically corresponds to the propagation delay for the transmission containing the timing information. If the user terminals and system both derive their timing from a common time source (e.g., GPS), then there may be no timing skews between these entities.

[0044] In FIG. 10A, user terminals A and B (e.g., randomly) select two different CH1 slots (e.g., slots 3 and 1, respectively) to transmit their CH1 PDUs. Because user terminals A and B are associated with different timing skews and different propagation delays, their CH1 PDUs arrive at the access point with different delays (referred to as round trip delays or RTDs) with respect to the access point's CH1 slot boundaries.

[0045] FIG. 10B shows an exemplary transmission on Channel 2. The time line for Channel 2 is divided into CH2 slots, with each CH2 slot having a particular duration (e.g., one OFDM symbol period). One CH2 PDU may be transmitted on each CH2 slot.

[0046] For FIG. 10B, user terminals A and B have locked their timing to that of the system and further have knowledge of their RTDs, which may be determined by the access point (e.g., during system access) and reported back to the user terminals. The user terminals may thereafter adjust their transmit timing to account for their RTDs such that their CH2 PDUs arrive time-aligned to the selected CH2 slot boundaries at the access point.

[0047] In FIG. 10B, user terminals A and B (e.g., randomly) select CH2 slots 3 and 1, respectively, to transmit their CH2 PDUs. Because user terminals A and B time-compensated their transmissions, the CH2 PDUs arrive at the access point

approximately aligned to the boundaries of the selected CH2 slots, as shown in FIG. 10B.

[0048] FIG. 4 shows a block diagram of an embodiment of a receiver unit 400 that can perform receive data processing for Channels 1 and 2 described above. Receiver unit 400, which may also be implemented within an access point or a user terminal, includes one receiver (RCVR) 410 for each of T receive antennas 408, a detector/data demodulator 420, and a receive (RX) data processor 450.

[0049] Each antenna 408 receives the RF modulated signals transmitted by the transmitter unit and provides a received signal to a respective receiver 410. Each receiver 410 conditions (e.g., amplifies, filters, and frequency downconverts) its received signal and digitizes the conditioned signal to provide samples, which are denoted as $x_i(n)$.

[0050] Detector/data demodulator 420 includes a data demodulator 430 and a detector 440 that receive and process the samples from all receivers 410 to detect and demodulate data transmissions on Channels 1 and 2. The processing by unit 420 is described in further detail below. Unit 420 provides recovered data symbols, denoted as $\hat{s}(k)$, which are estimates of the transmitted data symbols $s(k)$. Within RX data processor 450, the recovered data symbols are demapped by a symbol demapping unit 452, deinterleaved by a deinterleaver 454, and decoded by a decoder 456 to provide decoded data for CH1 and CH2 messages. If a recovered message includes a CRC value, then a CRC checker 458 checks the message with the CRC value to determine whether it was decoded correctly or in error.

[0051] FIG. 11A shows the receiver processing for Channel 1, which is not time-compensated. Referring back to FIG. 10A, even though the transmitter units attempt to transmit on specific CH1 slots, the CH1 transmissions are not time-compensated and the resultant behavior of Channel 1 is similar to that of an unslotted channel. In this case, referring back to FIG. 11A, the receiver unit can use a sliding correlation detector to detect for CH1 transmissions, each of which may be received starting at any sample period.

[0052] The correlation detector, which may operate in the time domain, slides through the entire time span in which CH1 PDUs may be received, one sample period at a time. A detection window indicates the time period in which samples for one CH1 PDU are to

be processed by the detector. This detection window may be initialized to the start of the first CH1 slot and would then slide forward one sample period at a time. For each sample period, which corresponds to a hypothesis, the correlation detector processes the samples within the detection window to determine a metric for a CH1 PDU hypothesized to have been received starting at that sample period. If the metric exceeds a CH1 threshold, then the CH1 PDU is further decoded to recover the CH1 message. The metric may relate to signal energy or some other parameter. The CH1 threshold may be fixed or adaptive (e.g., dynamically determined based on the samples within the detection window).

[0053] FIG. 5 shows a block diagram of a correlation detector 440a, which is one embodiment of detector 440 in FIG. 4. The samples $x_i(n)$ for each of the T receive antennas are provided to a respective antenna processor 510. Within each processor 510, a symbol accumulator 520 receives and accumulates the samples for the current hypothesis and provides accumulated samples $\tilde{x}_i(n)$ to a delay line/buffer 530. For the CH1 PDU shown in FIG. 2A, symbol accumulator 520 performs accumulation of the P pilot OFDM symbols, where the accumulation is performed on a per sample basis, to provide an accumulated pilot OFDM symbol having M samples. Delay line/buffer 530 provides storage for N of the M samples and effectively discards $\mathrm{M}-\mathrm{N}$ samples for the cyclic prefix. These N samples are for the transformed symbol corresponding to the accumulated pilot OFDM symbol,.

[0054] A signal detector 540 then determines a metric for the accumulated pilot OFDM symbol. In an embodiment and as described below, the metric relates to the signal energy of the N samples for the accumulated pilot OFDM symbol. However, other metrics may also be used, and this is within the scope of the invention. An adaptive threshold computation unit 550 determines an adaptive threshold value $Y_i(n)$ to use to decide whether or not a CH1 transmission was received. A summer 560 sums the threshold values for all T antennas to provide a combined threshold value $Y_{tot}(n)$, which is further scaled with a scaling factor S_1 by a multiplier 562 to obtain a final threshold value $Y(n)$. A summer 564 sums the metric values for all T antennas to provide a final metric value $E(n)$, which is then compared against the final threshold value $Y(n)$ by a

comparator 570. The detector output would indicate that a CH1 PDU was received if $E(n) > Y(n)$, and that no CH1 PDU was received otherwise.

[0055] FIG. 6 shows a block diagram of a correlation detector 440b, which is one embodiment of detector 440a in FIG. 5. The samples $x_i(n)$ for each receive antenna are provided to symbol accumulator 520, which is implemented with $\mathrm{P}-1$ delay units 522 and $\mathrm{P}-1$ summers 524. Each delay unit 522 provides one OFDM symbol (i.e., M samples) of delay. The $\mathrm{P}-1$ summers 524 perform accumulation of the P pilot OFDM symbols on a per sample basis, and the last summer provides the samples $\tilde{x}_i(n)$ for the accumulated pilot OFDM symbol. The samples $\tilde{x}_i(n)$ may be expressed as:

$$\tilde{x}_i(n) = \sum_{j=0}^{\mathrm{P}-1} x_i(n - j\mathrm{M}) \quad \text{, for } i \in \{1 \;...\; \mathrm{T}\}. \qquad \text{Eq (1)}$$

The samples $\tilde{x}_i(n)$ are provided to delay line/buffer 530, which is implemented with $\mathrm{N}-1$ delay units 532, each of which provides one sample period of delay.

[0056] Signal detector 540 performs correlation of the accumulated pilot OFDM symbol with the known pilot OFDM symbol and determines the metric value $E_i(n)$ for the accumulated pilot OFDM symbol. Each of the N samples for the accumulated pilot OFDM symbol is provided to a respective multiplier 542, which also receives a corresponding conjugated pilot sample $\tilde{p}_1^*(j)$, where $j \in \{0 \;...\; \mathrm{N}-1\}$. To obtain $\{\tilde{p}_1^*(j)\}$, the set of pilot symbols $\{p_1(k)\}$ for the pilot subbands and zero signal values for the unused subbands (e.g., as shown in Table 1) are transformed to the time domain using an N-point IFFT to obtain N pilot samples, $\tilde{p}_1(0)$ through $\tilde{p}_1(\mathrm{N}-1)$, which are then conjugated and provided to N multipliers 542. Each multiplier 542 multiplies its sample $\tilde{x}_i(n-j)$ with its conjugated pilot sample $\tilde{p}_1^*(j)$ and provides the result to a summer 544. Summer 544 sums the results from all N multipliers 542 and provides the summed result to a unit 546. Unit 546 determines the squared magnitude of the summed result, which is provided as the metric value $E_i(n)$. The metric value for each antenna may be expressed as:

$$E_i(n) = \left| \sum_{j=0}^{N-1} \tilde{p}_1^*(j) \cdot \tilde{x}_i(n-j) \right|^2 \quad , \text{ for } i \in \{1 \ldots \text{T}\} \ . \qquad \text{Eq (2)}$$

[0057] Summer 564 receives and sums the metric values for all T antennas to provide the final metric value $E(n)$, which may be expressed as:

$$E(n) = \sum_{i=1}^{\text{T}} E_i(n) \quad . \qquad \text{Eq (3)}$$

[0058] Threshold computation unit 550 determines an adaptive threshold to use for the detection of CH1 PDU for the current hypothesis. Each of the N samples for the accumulated pilot OFDM symbol is provided to a respective unit 552, which determines the squared magnitude of the sample. A summer 554 then sums the squared magnitudes from all N units 552 to provide the threshold value $Y_i(n)$. Summer 560 receives and sums the threshold values for all T antennas to provide the combined threshold value $Y_{tot}(n)$, which may be expressed as:

$$Y_{tot}(n) = \sum_{i=1}^{\text{T}} \sum_{j=0}^{N-1} \left| \tilde{x}_i(n-j) \right|^2 \quad . \qquad \text{Eq (4)}$$

Multiplier 562 then scales the combined threshold value with the scaling factor S_1 to provide the final threshold value, which may be given as $Y(n) = S_1 \cdot Y_{tot}(n)$.

[0059] Comparator 570 compares the final metric value $E(n)$ against the final threshold value $Y(n)$ and provides the detector output $D(n)$, which may be expressed as:

$$D(n) = \begin{cases} \text{"CH1 PDU present"} & \text{if } E(n) > Y(n) \ , \\ \text{"CH1 PDU not present"} & \text{otherwise} \quad . \end{cases} \qquad \text{Eq (5)}$$

If a CH1 PDU is detected, then the OFDM symbol timing is set at the time instant of the CH1 PDU detection (i.e., at the specific value of n when the CH1 PDU is detected).

[0060] The scaling factor S_1 is a positive constant selected to provide (1) a particular missed detection probability, which is the probability of not detecting a CH1 PDU that has been transmitted, and (2) a particular false alarm rate, which is the probability of

falsely indicating that a CH1 PDU was received when in fact none was transmitted. It is desirable to have the missed detection probability be less than the message error rate (MER), so that the MER is dictated by the received SNR and other parameters and not by the detector. The MER may be specified for Channel 1, for example, to be 1 percent or less. The detector output may be used to determine whether or not to process the received CH1 PDU to recover the transmitted CH1 message. The determination as to whether the CH1 message is decoded correctly or in error may be made based on a CRC value included in the message.

[0061] For a given received CH1 PDU, it may be possible for the correlation detector to declare multiple detections. This is because a detection may be declared with noise in one or more OFDM symbols and signal in the other OFDM symbols for the CH1 PDU being detected. For example, when $P = 2$, a first detection may occur with noise in OFDM symbol 1 and signal in OFDM symbol 2, and a second detection with a larger final metric value will occur when the second signal OFDM symbol arrives one OFDM symbol period later. Thus, for $P > 1$, the detector may be operated to continue to detect for the CH1 PDU for an additional $P-1$ OFDM symbol periods to find the largest final metric value for the PDU. The OFDM symbol timing is then set by the detection with the largest final metric value and the RTD is also computed based on the time associated with this detection.

[0062] The detection processing may be performed independently of the message processing, i.e., the detection processing can continue in the normal manner regardless of whether or not CH1 PDUs are detected. Thus, if a CH1 PDU is initially detected at sample period $n-j$ with a final metric value of $E(n-j)$ and another CH1 PDU is later detected at sample period n with a final metric value of $E(n)$, where $E(n) > E(n-j)$ and j is smaller than the size of the detection window, then the current message processing for the CH1 PDU detected at sample period $n-j$ may be halted and the CH1 PDU detected at sample period n may be processed instead.

[0063] FIG. 11B shows the receiver processing for Channel 2, which is time-compensated. Referring back to FIG. 10B, the transmitter units transmit on specific CH2 slots and the CH2 transmissions are time-compensated to arrive at the receiver unit at the selected CH2 slot boundaries. In this case, referring back to FIG. 11B, the receiver unit can detect for CH2 transmissions in each CH2 slot (instead of each sample

period), and the detection window can move from slot to slot. For each CH2 slot, which corresponds to a hypothesis, the decision-directed detector processes the samples received within the detection window to determine a metric for a CH2 PDU hypothesized to have been received in that slot. If the metric exceeds a CH2 threshold, then the CH2 PDU is deemed to have been received.

[0064] FIG. 7 shows a block diagram of an embodiment of a detector/data demodulator 420c, which may also be used for unit 420 in FIG. 4. Detector/data demodulator 420c includes a data demodulator 430c used to perform coherent demodulation and a decision-directed detector 440c used to detect for CH2 PDUs. The samples for each of the T receive antennas are provided to a respective antenna demodulator 710 within data demodulator 430c and to a respective decision-directed detector 750 within detector 440c.

[0065] Each antenna demodulator 710 performs coherent demodulation for one antenna for one received OFDM symbol at a time. For each received OFDM symbol, an FFT unit 712 receives the samples $x_i(n)$ for the OFDM symbol, removes the cyclic prefix to obtain the transformed symbol, and performs a fast Fourier transform (FFT) on the transformed symbol to provide N received symbols $r_i(k)$, which include received data symbols $r_{i,d}(k)$ and received pilot symbols $r_{i,p}(k)$. A channel estimator 720 then estimates the channel response of the data subbands based on the received pilot symbols $r_{i,p}(k)$. A demodulator 730 performs coherent demodulation of the received data symbols $r_{i,d}(k)$ with the channel estimates to provide recovered data symbols $\hat{s}_i(k)$.

[0066] A symbol accumulator 740 receives and accumulates the recovered data symbols from demodulators 710a through 710t for the T receive antennas and provides recovered symbols $\hat{s}(k)$. RX data processor 450 then processes the recovered symbols $\hat{s}(k)$, as described above for FIG. 4, to provide the decoded data. In an embodiment, the CH2 message does not include a CRC, and the CRC check is not performed by the RX data processor. A TX data processor 310 then processes the decoded data to provide remodulated symbols $c(k)$, which are estimates of the transmitted data symbols $s(k)$. The processing by processor 310 includes encoding, interleaving, and symbol mapping, as described above for FIG. 3A. The processing by RX data processor 450 is often

referred to as simply "decoding", and the processing by TX data processor 310 is often referred to as "re-encoding".

[0067] Each decision-directed detector 750 performs detection for one received OFDM symbol at a time. For each received OFDM symbol, an FFT unit 752 receives the samples $x_i(n)$ for the OFDM symbol and performs an FFT on the corresponding transformed symbol to provide N received symbols $r_i(k)$. FFT units 712 and 752 are typically implemented with one FFT unit, but are shown as two units in FIG. 7 for clarity.

[0068] A signal detector 760 then processes the received pilot and data symbols with their expected symbols to provide a metric $E_i'(n)$ for the OFDM symbol being processed. An adaptive threshold computation unit 770 determines an adaptive threshold value $Y_i'(n)$ used to decide whether or not a CH2 PDU was received. A summer 780 sums the threshold values for all T antennas to provide a combined threshold value $Y_{tot}'(n)$, which is further scaled with a scaling factor S_2 by a multiplier 782 to obtain a final threshold value $Y'(n)$. A summer 784 sums the metric values for all T antennas to provide the final metric value $E'(n)$, which is then compared against the final threshold value $Y'(n)$ by a comparator 790. The detector output would indicate that a CH2 PDU was received if $E'(n) > Y'(n)$, and that no CH2 PDU was received otherwise.

[0069] FIG. 8A shows a block diagram of a data demodulator 430d, which is one embodiment of data demodulator 430c in FIG. 7. The samples $x_i(n)$ for each receive antenna are transformed by FFT unit 712 to provide N received symbols $r_i(k)$ for each transformed symbol. For the embodiment shown in Table 1, the N received symbols include 28 received pilot symbols for 28 pilot subbands, 24 received data symbols for 24 data subbands, and 12 additional symbols for the 12 unused subbands. For simplicity, the following description is for the embodiment shown in FIG. 2B whereby the N received symbols include $L+1$ received pilot symbols for $L+1$ pilot subbands and L received data symbols for L data subbands, where each data subband is flanked

on both sides by pilot subbands, and the subband index k for the pilot and data subbands is defined as $k \in K$ where $K = \{1 \ldots 49\}$.

[0070] Coherent demodulation of each of the L data subbands is performed by first forming an estimate of the channel response for the data subband using the two pilot subbands flanking the data subband. The channel estimate $\hat{h}_i(k)$ for the k-th data subband may be obtained by combining the channel estimates for the two flanking pilot subbands, which may be expressed as:

$$\begin{aligned}\hat{h}_i(k) &= \hat{h}_i(k-1) + \hat{h}_i(k+1) \\ &= p_2^*(k-1)r_i(k-1) + p_2^*(k+1)r_i(k+1)\end{aligned} \quad \text{, for } k \in K_d \text{ and } i \in \{1 \ldots \mathrm{T}\}, \quad \text{Eq (6)}$$

where $p_2(k)$ is the pilot symbol transmitted on the k-th subband for Channel 2 and K_d represents the set of data subbands, i.e., $K_d \in \{2, 4, \ldots\ 2\mathrm{L}\}$.

[0071] The recovered data symbol $\hat{s}_i(k)$ for each data subband may then be expressed as:

$$\hat{s}_i(k) = \hat{h}_i^*(k) \cdot r_i(k) \text{ , for } k \in K_d \text{ and } i \in \{1 \ldots \mathrm{T}\}. \qquad \text{Eq (7)}$$

The recovered data symbols for all T receive antennas for each data subband may then be obtained as:

$$\hat{s}(k) = \sum_{i=1}^{\mathrm{T}} \hat{s}_i(k) \text{ , for } k \in K_d. \qquad \text{Eq (8)}$$

[0072] In FIG. 8A, the channel estimation shown in equation (6) is performed by $\mathrm{L}+1$ multipliers 722 and L summers 724. Each multiplier 722 multiplies the received symbol for a respective pilot subband with the conjugate of the known pilot symbol for that subband to provide the channel estimate for the pilot subband. Each summer 724 then sums the channel estimates for the two pilot subbands flanking the associated data subband to provide the channel estimate for that data subband. The channel estimates for the L data subbands may also be obtained based on interpolation or some other manners, and this is within the scope of the invention.

[0073] The coherent demodulation shown in equation (7) is performed by L multipliers 732. Each multiplier 732 multiplies the received symbol $r_i(k)$ for a respective data subband with the conjugate of the channel estimate, $\hat{h}_i^*(k)$, for that subband to provide the recovered data symbol $\hat{s}_i(k)$ for the data subband. Sample accumulation for all T receive antennas, as shown in equation (8), is performed by L summers 742. Each summer 742 receives and sums T recovered data symbols $\hat{s}_i(k)$ for the T receive antennas for the associated data subband to provide the recovered symbol $\hat{s}(k)$ for that subband.

[0074] As noted above, the subband multiplexing may be such that each group of Q data subbands is flanked on both sides by pilot subbands, where Q may be greater than one. If $Q > 1$, then coherent demodulation may be performed in several manners. In one embodiment, the received pilot symbol for each pilot subband is used as a coherent reference for the two adjacent data subbands, and the received data symbols for these data subbands may be coherently demodulated based on this received pilot symbol. Hard decisions may then be obtained and used to remove the modulation from the just-detected data symbols to obtain improved channel estimates for the next two data subbands. The demodulation process can start from the end data subbands (i.e., next to the pilot subbands) and work towards the middle data subband. Improved channel estimates for the data subbands further away from the pilot subbands may be obtained as each pair of received data symbols is detected. In another embodiment, the received pilot symbols for each pair of pilot subbands are interpolated to obtain the channel estimate for each of the Q data subbands flanked by these pilot subbands.

[0075] A CRC value is often used to determine whether a received message was decoded correctly or in error. In certain instances, it may not be desirable to include a CRC value in a message because of the overhead associated with the CRC value and/or some other consideration. In this case, another mechanism is needed to determine whether or not the received message is valid. For the embodiment shown in FIG. 7, data demodulator 430c and RX data processor 450 may be operated to provide a decoded message for each hypothesis, and detector 440c may be operated to provide an indication as to whether or not a message was received for the hypothesis.

[0076] FIG. 8B shows a block diagram of a decision-directed detector 440d that performs differential detection in the frequency domain and is one embodiment of detector 440c in FIG. 7. The samples $x_i(n)$ for each receive antenna are transformed by FFT unit 752 to provide N received symbols $r_i(k)$ for each transformed symbol.

[0077] To determine the metric value $E'(n)$ for each transformed symbol, a detection statistic $g_i(n)$ is first obtained for each receive antenna by summing over the real part of 2L dot products formed by using adjacent pairs of pilot and data subbands. The detection statistic $g_i(n)$ may be expressed as:

$$g_i(n) = \sum_{k=1}^{2L} z(k) \cdot z^*(k+1) \quad \text{for } i \in \{1 \ldots \mathrm{T}\} \ , \qquad \text{Eq (9a)}$$

where

$$z_i(k) = \begin{cases} r_i(k) \cdot p_2^*(k) & \text{for } k \in \{1,\ 3,\ \ldots\ 2\mathrm{L}+1\} \\ r_i(k) \cdot c^*(k) & \text{for } k \in \{2,\ 4,\ \ldots\ 2\mathrm{L}\} \end{cases} . \qquad \text{Eq (9b)}$$

[0078] The metric value $E'(n)$ for the transformed symbol may then be expressed as:

$$E'(n) = \left| \sum_{i=1}^{\mathrm{T}} \mathrm{Re}\{g_i(n)\} \right|^2 \quad \text{for } i \in \{1 \ldots \mathrm{T}\} \ . \qquad \text{Eq (10a)}$$

Alternatively, the metric value $E'(n)$ may be expressed as:

$$E'(n) = \sum_{i=1}^{\mathrm{T}} \left| \mathrm{Re}\{g_i(n)\} \right|^2 \quad \text{for } i \in \{1 \ldots \mathrm{T}\} \ . \qquad \text{Eq (10b)}$$

[0079] In FIG. 8B, the computation of the detection statistic $g_i(n)$ shown in equation (9) is performed by $2\mathrm{L}+1$ multipliers 762, $2\mathrm{L}$ multipliers 764, and a summer 766. Each multiplier 762 multiplies the received symbol for an associated pilot or data subband with the conjugate of the known pilot symbol or remodulated symbol for that subband. Each multiplier 764 performs a dot product of the outputs from a pair of multipliers 762 for a pair of adjacent pilot and data subbands. Summer 766 then sums the outputs from L multipliers 764 to provide the detection statistic $g_i(n)$. For the

embodiment shown in equation (10a), a unit 768 receives $g_i(n)$ and provides the real part to summer 784, which sums the real part of $g_i(n)$ for all T antennas. The output from summer 784 is then squared by a unit 786 to provide the metric value $E'(n)$. For the embodiment shown in equation (10b), unit 786 may be placed between unit 768 and summer 784.

[0080] Adaptive threshold computation unit 770 determines the adaptive threshold $Y'(n)$ to use for each received transformed symbol. Each of the $2L+1$ received symbols $r_i(k)$ for the pilot and data subbands is provided to a respective unit 772, which determines the squared magnitude of the symbol. A summer 774 then sums the squared magnitude from all $2L+1$ units 772 to provide the threshold value $Y_i'(n)$. Summer 780 receives and sums the threshold values for all T antennas to provide the combined threshold value $Y_{tot}'(n)$, which may be expressed as:

$$Y_{tot}'(n) = \sum_{i=1}^{T} \sum_{k=1}^{2L+1} |r_i(k)|^2 \quad . \qquad \text{Eq (11)}$$

Multiplier 782 scales the combined threshold value with a scaling factor S_{2a} to provide the final threshold value, which may be given as $Y'(n) = S_{2a} \cdot Y_{tot}'(n)$. In general, the threshold value $Y'(n)$ and metric value $E'(n)$ are each accumulated over the duration of the PDU to be detected. Thus, if the PDU spans multiple OFDM symbol periods, then the threshold and metric values are first computed as described above for each of these OFDM symbols and then accumulated to provide the final threshold and metric values for the PDU.

[0081] Comparator 790 compares the final metric value $E'(n)$ against the final threshold value $Y'(n)$ and provides the detector output $D'(n)$, which may be expressed as:

$$D'(n) = \begin{cases} \text{"CH2 PDU present"} & \text{if } E'(n) > Y'(n) \;, \\ \text{"CH2 PDU not present"} & \text{otherwise} \qquad . \end{cases} \qquad \text{Eq (12)}$$

If the detector output $D'(n)$ indicates that a CH2 PDU is present, then the CH2 message decoded by the RX data processor is deemed to be valid and may be further processed by a controller as appropriate. Otherwise, the CH2 message is discarded.

[0082] FIG. 8C shows a block diagram of a decision-directed detector 440e that performs coherent detection in the time domain and is another embodiment of detector 440c in FIG. 7. The samples $x_i(n)$ for each receive antenna are provided to a delay line/buffer 830 that is implemented with $\mathrm{N}-1$ delay units 832, each of which provides one sample period of delay.

[0083] Detector 440e performs correlation of each received OFDM symbol with its corresponding "reconstructed" OFDM symbol to determine the metric $E''(n)$ for the received OFDM symbol. Each of the N samples $x_i(n)$ for the received OFDM symbol is provided to a respective multiplier 842, which also receives a corresponding conjugated reconstructed sample $d^*(j)$, where $j \in \{0 \ldots \mathrm{N}-1\}$. To obtain $d^*(j)$, the pilot symbols $p_2(k)$ for the pilot subbands (e.g., as shown in Table 1), remodulated symbols $c(k)$ for the data subbands, and zero signal values for the unused subbands (i.e., N symbols for the N total subbands) for an OFDM symbol period are transformed to the time domain by an N-point IFFT 830 to obtain N reconstructed samples, $d(0)$ through $d(\mathrm{N}-1)$, which are then conjugated and provided to N multipliers 842. The operations performed by the other elements in FIG. 8C are as described above for FIG. 6. The metric value $E_i''(n)$ for each antenna may be expressed as:

$$E_i''(n) = \left| \sum_{j=0}^{\mathrm{N}-1} d^*(j) \cdot x_i(n-j) \right|^2 \quad , \text{ for } i \in \{1 \ldots \mathrm{T}\} \ . \qquad \text{Eq (13)}$$

The final metric value $E''(n)$ for all T antennas may then be expressed as:

$$E''(n) = \sum_{i=1}^{\mathrm{T}} E_i''(n) \quad . \qquad \text{Eq (14)}$$

[0084] The threshold $Y''(n)$ to use for comparing against the final metric value $E''(n)$ may be determined as described above for FIG. 6. In particular, the combined threshold value $Y_{tot}''(n)$ for all T antennas may be expressed as:

$$Y''_{tot}(n) = \sum_{i=1}^{T}\sum_{j=0}^{N-1}|x_i(n-j)|^2 \quad . \qquad \text{Eq (15)}$$

The final threshold value may then be given as $Y''(n) = S_{2b} \cdot Y''_{tot}(n)$.

[0085] For the decision-directed detector, the scaling factor S_2 (which is S_{2a} for detector 440d in FIG. 8B and S_{2b} for detector 440e in FIG. 8C) is a positive constant selected to provide (1) a particular missed detection probability for CH2 PDUs and (2) a particular false alarm rate for incorrectly declaring the presence of CH2 PDUs. If CH2 messages are defined such that they do not include CRC values, then the detector is relied upon exclusively to determine whether or not CH2 messages are present. Erroneous CH2 messages may be provided to the controller due to the following:

- false alarm - noise in the received signal falsely triggers detection; and
- incorrect decode - signal correctly triggers detection but the decoded CH2 message includes uncorrected and undetected errors.

[0086] If Channel 2 is used as a random access channel, then a false alarm for a CH2 PDU may cause the system to assign resources to a non-existent user terminal, which then results in wasted resources. In that case, it is desirable to select the scaling factor S_2 to minimize the false alarm probability since it is undesirable to have noise frequently triggering a waste of resources.

[0087] The incorrect decode probability is related to the detection probability, and a higher detection probability can lead to more incorrect decode events. When an incorrect decode event occurs, an erroneously decoded CH2 message is provided to the controller. The controller may be able to check the validity of the CH2 message in some other manner. For example, if the CH2 message includes a unique identifier for the user terminal that transmitted the message, then the controller can check to see if the unique identifier for the recovered CH2 message is included in a list of valid identifiers. If the unique identifier in the received CH2 message is determined to be valid, then the system can assign resources to the user terminal associated with that identifier.

[0088] In selecting the scaling factor S_2, it may be desirable to detect as many valid CH2 messages as possible while maintaining the false alarm rate and incorrect decode probability to below a particular level. It is also possible to vary the scaling factor S_2

based on system loading. For example, if the system load is low and there are few valid identifiers, then the likelihood of the system erroneously allocating resources is smaller. In this case, a lower detection threshold may be used. As the system load increases, the detection threshold may be increased to reduce the rate of incorrect decode events.

[0089] FIG. 9 shows a block diagram of an embodiment of an access point 110x and a user terminal 120x in system 100. For this embodiment, access point 110x and user terminal 120x are each equipped with multiple antennas. In general, the access point and user terminal may each be equipped with any number of transmit/receive antennas.

[0090] On the uplink, at user terminal 120x, TX data processor 310 receives and processes traffic data from a data source 308 and other data (e.g., for CH1 and CH2 messages) from a controller 360 to provide multiplexed data and pilot symbols, as described above for FIG. 3A. TX spatial processor 320 may perform spatial processing on the pilot and data symbols to provide a stream of transmit symbols for each antenna. Each modulator 340 receives and processes a respective transmit symbol stream to provide a corresponding uplink modulated signal, which is then transmitted from an associated antenna 350.

[0091] At access point 110x, T antennas 408a through 408t receive the transmitted uplink modulated signals from the user terminal, and each antenna provides a received signal to a respective receiver 410. Each receiver 410 conditions the received signal and further digitizes the conditioned signal to provide samples. Detector/data demodulator 420 then performs processing to detect for CH1 and CH2 messages, as described above. RX data processor 450 processes recovered symbols to provide decoded traffic data (which may be provided to a data sink 452 for storage) and recovered CH1 and CH2 messages (which may be provided to a controller 460 for further processing).

[0092] The processing for the downlink may be the same or different from the processing for the uplink. Data from a data source 468 and signaling (e.g., reply messages) from controller 460 are processed (e.g., coded, interleaved, and modulated) by a TX data processor 470 and may be spatially processed by a TX spatial processor 480. The transmit symbols from TX spatial processor 480 are then processed by modulators 410a through 410t to generate T downlink modulated signals, which are transmitted via antennas 408a through 408t.

[0093] At user terminal 120x, the downlink modulated signals are received by antennas 350, conditioned and digitized by receivers 340, and processed by an RX spatial processor 370 and an RX data processor 380 in a complementary manner to that performed at the access point. The decoded data for the downlink may be provided to a data sink 382 for storage and/or controller 360 for further processing.

[0094] Controllers 360 and 460 control the operation of various processing units at the user terminal and the access point, respectively. Memory units 362 and 462 store data and program codes used by controllers 360 and 460, respectively.

[0095] For clarity, specific embodiments of the correlation and decision-directed detectors, demodulators, and the receiver units have been described for specific PDU formats. Various other embodiments and uses for these detectors are also possible, and this is within the scope of the invention. For example, the correlation detector may be used for a channel whereby transmissions are time-compensated, and the decision-directed detector may be used for a channel whereby transmissions are not time-compensated.

[0096] The decision-directed detector may be implemented in the frequency domain (as shown in FIG. 8B) or the time domain (as shown in FIG. 8C). Moreover, the decision-directed detector may be used for various PDU formats. For example, the decision-directed detector may be used for a PDU format whereby data and pilot are subband multiplexed (as described above for CH2 PDU), a PDU format whereby data and pilot are time division multiplexed (TDM) (as described above for CH1 PDU), and others. The decision-directed detector may also be used with or without the pilot. In general, the decision-directed detector uses frequency-domain received data symbols or time-domain reconstructed data samples to detect for data transmissions in the received signal. This detector may advantageously be used when CRC or other error detection mechanisms are not available to detect for message errors.

[0097] The use of an adaptive threshold can provide robust detection performance in many operating scenarios, such as for an unlicensed frequency band where various sources of interference may be present. The threshold may be set based on a particular statistic for the transmission to be detected. This statistic may relate to the energy of the desired signal plus noise and interference in the transmission or some other parameter.

[0098] The detectors, demodulators, and receivers described herein may be used for various types of transport channels. For example, these units may be used for different

types of random access channels, such as the ones described in detail in the aforementioned U.S. Patent Application Serial No. 60/432,440 and provisional U.S. Patent Application Serial No. 60/421,309.

[0100] The detectors, demodulators, and receivers described herein may also be used for various wireless multiple-access communication systems. One such system is a wireless multiple-access MIMO system described in the aforementioned provisional U.S. Patent Application Serial No. 60/421,309. In general, these systems may or may not employ OFDM, or may employ some other multi-carrier modulation scheme instead of OFDM, and may or may not utilize MIMO.

[0101] The detectors, demodulators, and receivers described herein may be implemented by various means. For example, these units may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the detectors and receivers may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[0102] For a software implementation, the signal processing for the detectors, demodulators, and receivers may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory units 362 and 462 in FIG. 9) and executed by a processor (e.g., controllers 360 and 460). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[0103] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. A receiver unit in a wireless communication system, comprising:

a demodulator operative to process received data symbols to provide recovered symbols;

a first data processor operative to process the recovered symbols to provide decoded data;

a second data processor operative to process the decoded data to provide remodulated symbols; and

a detector operative to process the received data symbols and the remodulated symbols to provide a detector output.

2. The receiver unit of claim 1, wherein the received data symbols are for a data transmission hypothesized to have been received, and wherein the detector output indicates whether or not the data transmission is deemed to have been received.

3. The receiver unit of claim 2, further comprising:

a threshold computation unit operative to determine a threshold to use for the data transmission, and

wherein the detector is operative to provide a metric based on the received data symbols and the remodulated symbols, and wherein the detector output is determined based on the metric and the threshold.

4. The receiver unit of claim 3, wherein the threshold computation unit is operative to determine the threshold based on a plurality of received signals for a plurality of antennas, and wherein the detector is operative to determine the metric based on the plurality of received signals.

5. The receiver unit of claim 1, wherein the detector is further operative to process received pilot symbols to provide the detector output.

6. The receiver unit of claim 5, wherein data symbols are transmitted on data subbands and pilot symbols are transmitted on pilot subbands, and wherein the data subbands are multiplexed with the pilot subbands.

7. The receiver unit of claim 6, wherein the data subbands are interlaced with the pilot subbands such that each of the data subbands is flanked on both sides by pilot subbands.

8. The receiver unit of claim 1, wherein the detector is operative to perform coherent detection in time domain.

9. The receiver unit of claim 8, wherein the received data symbols are obtained based on input samples for a data transmission hypothesized to have been received, and wherein the detector is operative to perform correlation between the input samples and reconstructed samples obtained based on the remodulated symbols.

10. The receiver unit of claim 8, wherein the received data symbols are obtained based on input samples for a data transmission hypothesized to have been received, and wherein the detector is operative to perform correlation between the input samples and reconstructed samples obtained based on the remodulated symbols and pilot symbols for the data transmission.

11. The receiver unit of claim 1, wherein the detector is operative to perform differential detection in the frequency domain.

12. The receiver unit of claim 5, wherein the detector is operative to

multiply each of the received data symbols with a corresponding one of the remodulated symbols to provide a demodulated data symbol,

multiply each of the received pilot symbols with a corresponding one of known pilot symbols to provide a demodulated pilot symbol,

perform dot products between demodulated data symbols and demodulated pilot symbols, and

accumulate results of the dot products.

13. The receiver unit of claim 2, wherein the data transmission is for a random access channel in the wireless communication system.

14. The receiver unit of claim 1, wherein the wireless communication system uses multi-carrier modulation.

15. The receiver unit of claim 1, wherein the wireless communication system uses orthogonal frequency division multiplexing (OFDM).

16. A receiver unit in a wireless communication system, comprising:

a processor operative to process received data symbols for a data transmission hypothesized to have been received and provide remodulated symbols that are estimates of transmitted data symbols; and

a detector operative to process the received data symbols and the remodulated symbols to provide a detector output that indicates whether or not the data transmission is deemed to have been received.

17. The receiver unit of claim 16, wherein the processor is operative to demodulate the received data symbols to provide recovered symbols, decode the recovered symbols to provide decoded data, and re-encode the decoded data to provide the remodulated symbols.

18. The receiver unit of claim 16, wherein the processor is further operative to process received pilot symbols for the data transmission and corresponding known pilot symbols to provide the detector output.

19. A receiver unit in a wireless communication system, comprising:

a signal detector operative to determine a metric for a data transmission hypothesized to have been received;

a threshold computation unit operative to determine a threshold for the hypothesized data transmission; and

a comparator operative to receive the metric and the threshold and provide an output indicating whether or not the data transmission is deemed to have been received.

20. The receiver unit of claim 19, wherein the threshold is determined based on received pilot symbols for the hypothesized data transmission.

21. The receiver unit of claim 20, wherein the threshold is further determined based on received data symbols for the hypothesized data transmission.

22. The receiver unit of claim 19, wherein the metric relates to signal energy of the hypothesized data transmission.

23. The receiver unit of claim 19, wherein the signal detector is operative to determine the metric based on a plurality of received signals for a plurality of antennas, and wherein the threshold computation unit is operative to determine the threshold based on the plurality of received signals.

24. A method of detecting data transmissions in a wireless multiple-access communication system, comprising:

first processing received data symbols for a data transmission hypothesized to have been received to provide remodulated symbols that are estimates of transmitted data symbols; and

second processing the received data symbols and the remodulated symbols to provide a detector output that indicates whether or not the data transmission is deemed to have been received.

25. The method of claim 24, wherein the first processing includes demodulating the received data symbols to provide recovered symbols, decoding the recovered symbols to provide decoded data, and re-encoding the decoded data to provide the remodulated symbols.

26. The method of claim 24, further comprising:

determining a threshold to use for the hypothesized data transmission, and wherein the detector output is further determined based on the threshold.

27. The method of claim 26, wherein the second processing includes

determining a metric based on the received data symbols and the remodulated symbols, and

comparing the metric against the threshold, and wherein the detector output is based on the comparing.

28. A method of detecting data transmissions in a wireless multiple-access communication system, comprising:

determining a metric for a data transmission hypothesized to have been received;

determining a threshold for the hypothesized data transmission based on samples received for the hypothesized data transmission; and

comparing the metric against the threshold to provide an output indicating whether or not the data transmission is deemed to have been received.

29. An apparatus in a wireless multiple-access communication system, comprising:

means for processing received data symbols for a data transmission hypothesized to have been received to provide remodulated symbols that are estimates of transmitted data symbols; and

means for processing the received data symbols and the remodulated symbols to provide a detector output that indicates whether or not the data transmission is deemed to have been received.

30. The apparatus of claim 29, further comprising:

means for demodulating the received data symbols to provide recovered symbols;

means for decoding the recovered symbols to provide decoded data; and

means for re-encoding the decoded data to provide the remodulated symbols.

31. An apparatus in a wireless multiple-access communication system, comprising:

means for determining a metric for a data transmission hypothesized to have been received;

means for determining a threshold for the hypothesized data transmission based on samples received for the hypothesized data transmission; and

means for comparing the metric against the threshold to provide an output indicating whether or not the data transmission is deemed to have been received.

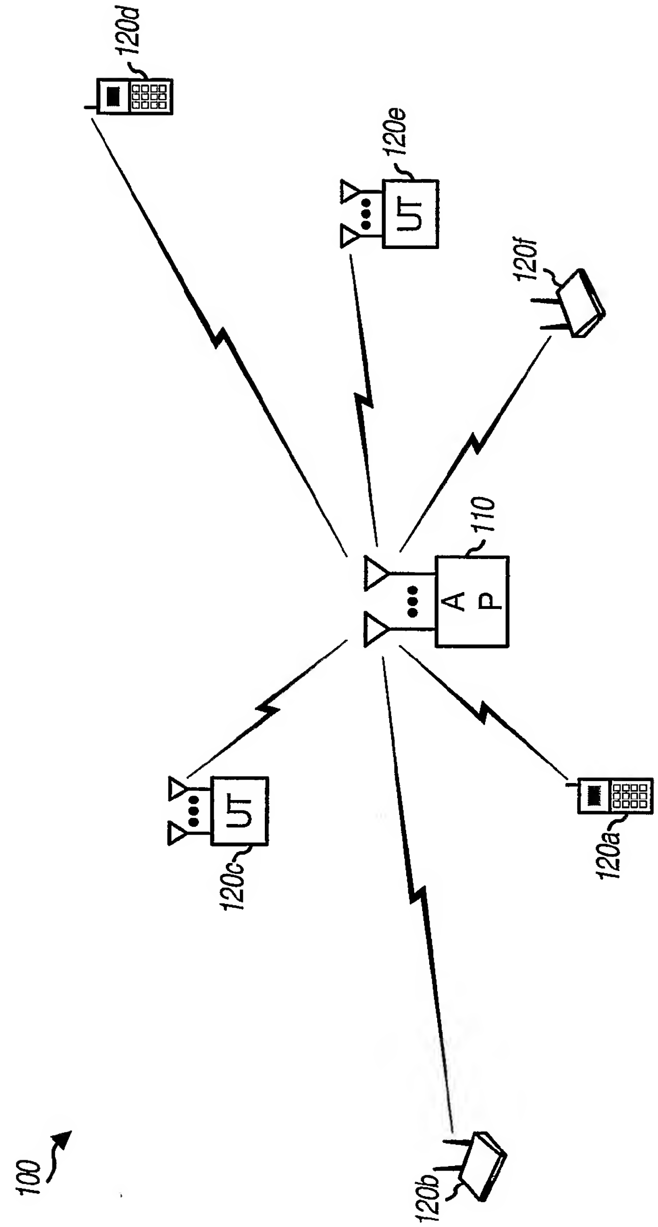


FIG. 1

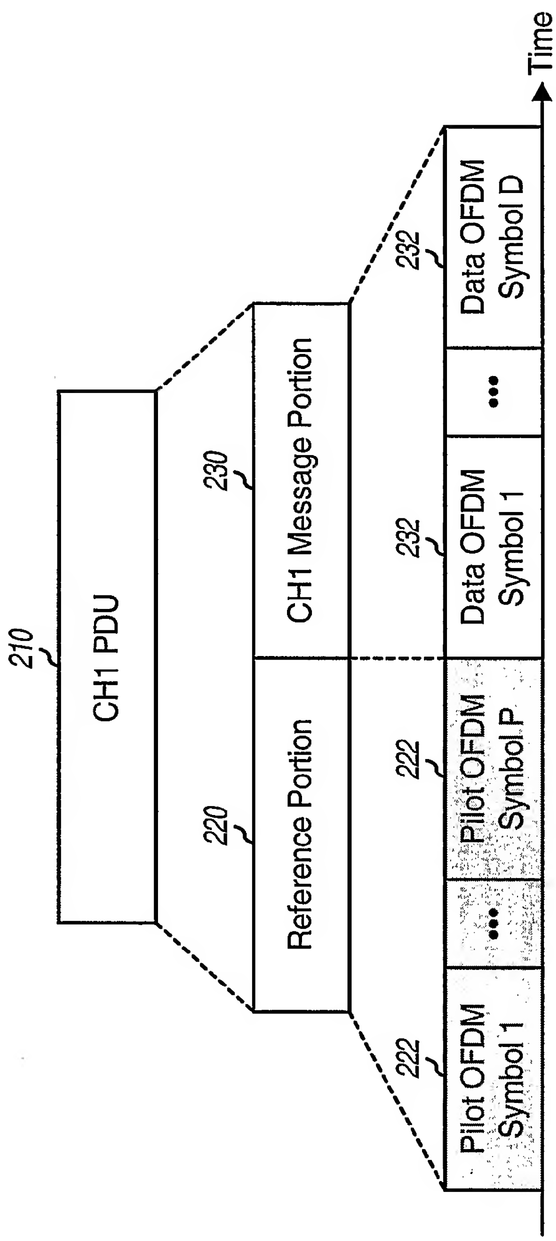
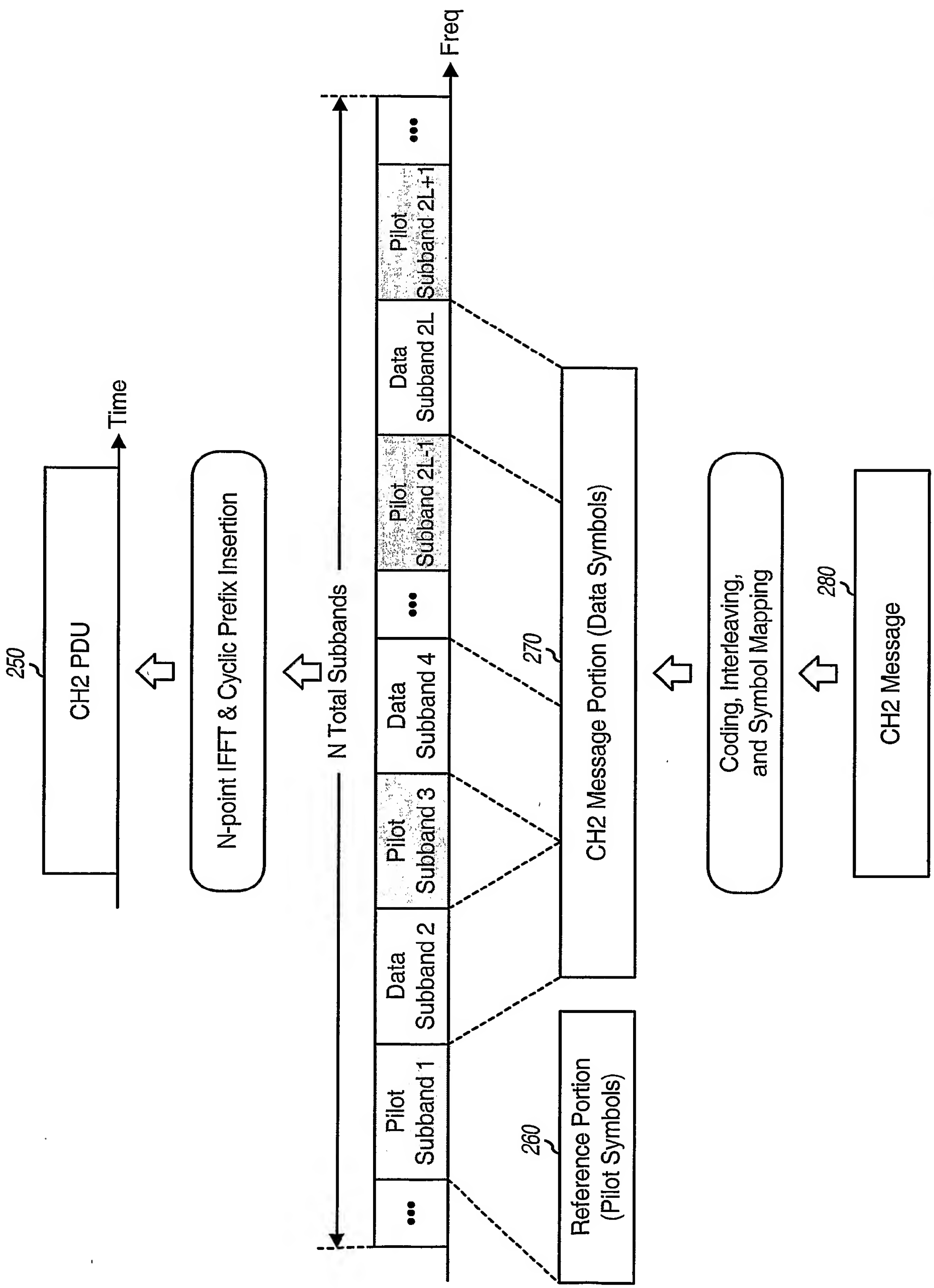

210
CH1 PDU
220
Reference Portion
230
CH1 Message Portion
222
Pilot OFDM Symbol 1
222
Pilot OFDM Symbol P
232
Data OFDM Symbol 1
232
Data OFDM Symbol D
Time

FIG. 2A



250
CH2 PDU
Time
N-point IFFT & Cyclic Prefix Insertion
N Total Subbands
Freq
Pilot Subband 1
Data Subband 2
Pilot Subband 3
Data Subband 4
Pilot Subband 2L-1
Data Subband 2L
Pilot Subband 2L+1
260
Reference Portion (Pilot Symbols)
270
CH2 Message Portion (Data Symbols)
Coding, Interleaving, and Symbol Mapping
280
CH2 Message

FIG. 2B

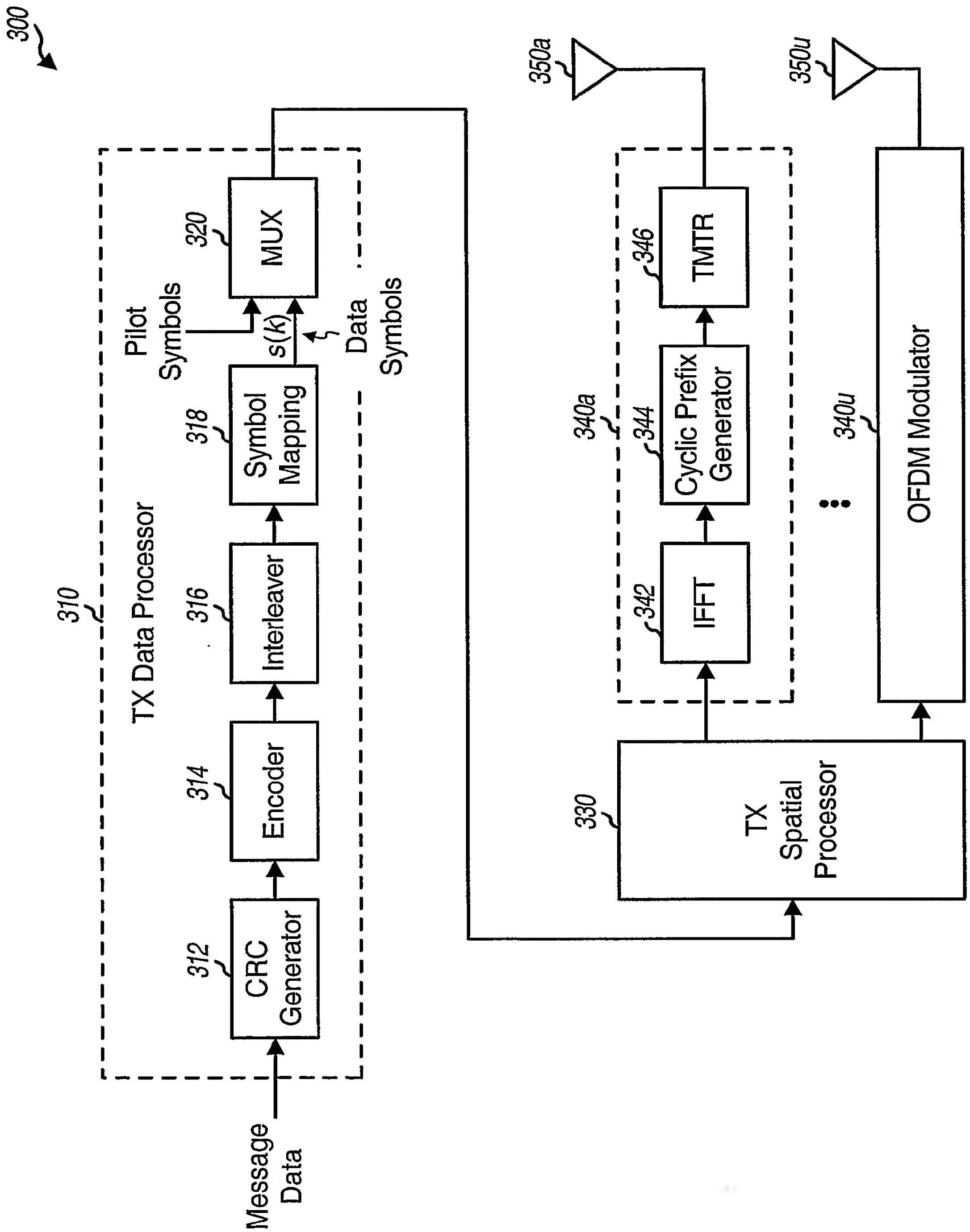

300
310
TX Data Processor
Message Data
312
CRC Generator
314
Encoder
316
Interleaver
318
Symbol Mapping
s(k)
Data Symbols
Pilot Symbols
320
MUX
330
TX Spatial Processor
340a
342
IFFT
344
Cyclic Prefix Generator
346
TMTR
350a
340u
OFDM Modulator
350u

FIG. 3A

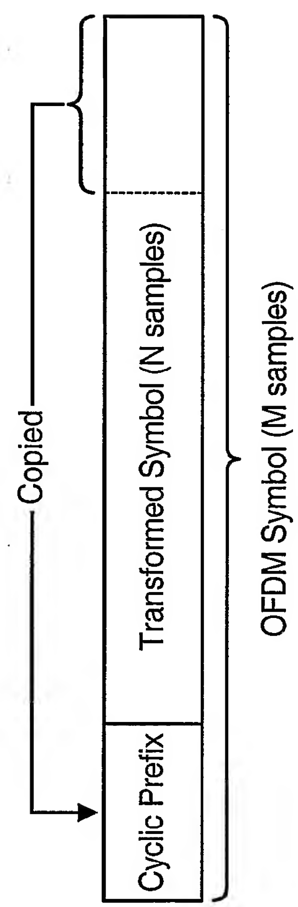

Copied
Cyclic Prefix
Transformed Symbol (N samples)
OFDM Symbol (M samples)

FIG. 3B

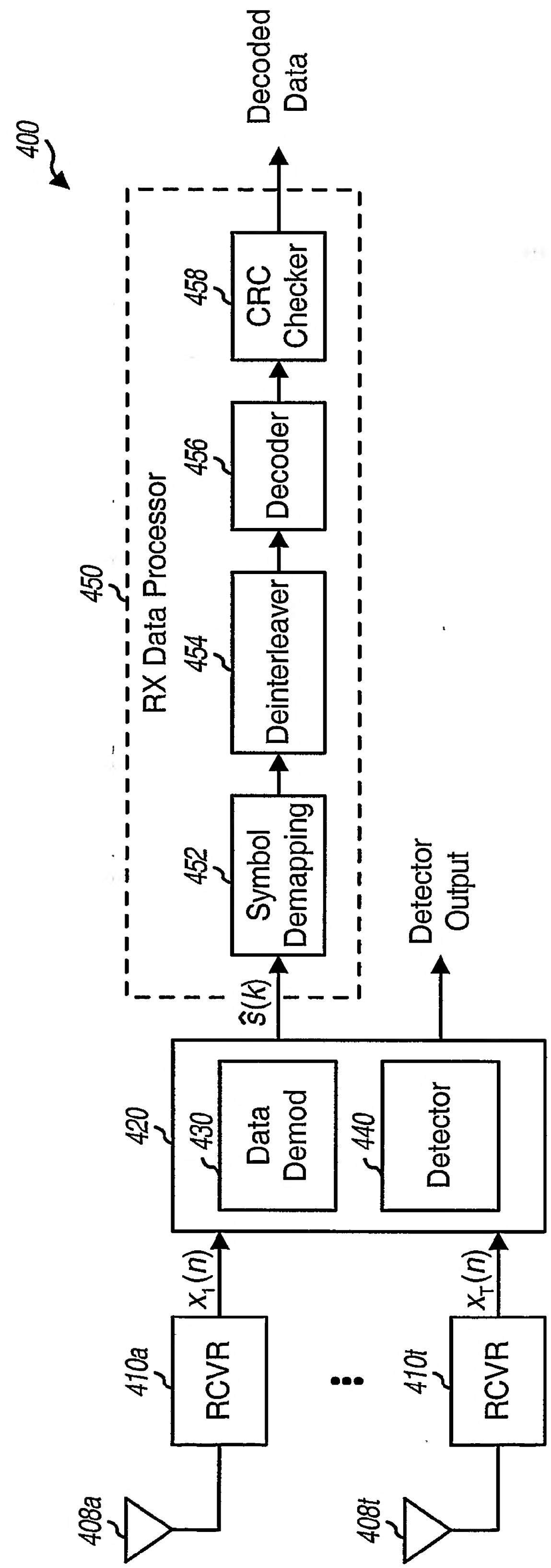

400
408a
408t
410a
410t
RCVR
RCVR
$x_1(n)$
$x_T(n)$
420
430
Data Demod
440
Detector
Detector Output
$\hat{s}(k)$
450
RX Data Processor
452
Symbol Demapping
454
Deinterleaver
456
Decoder
458
CRC Checker
Decoded Data

FIG. 4

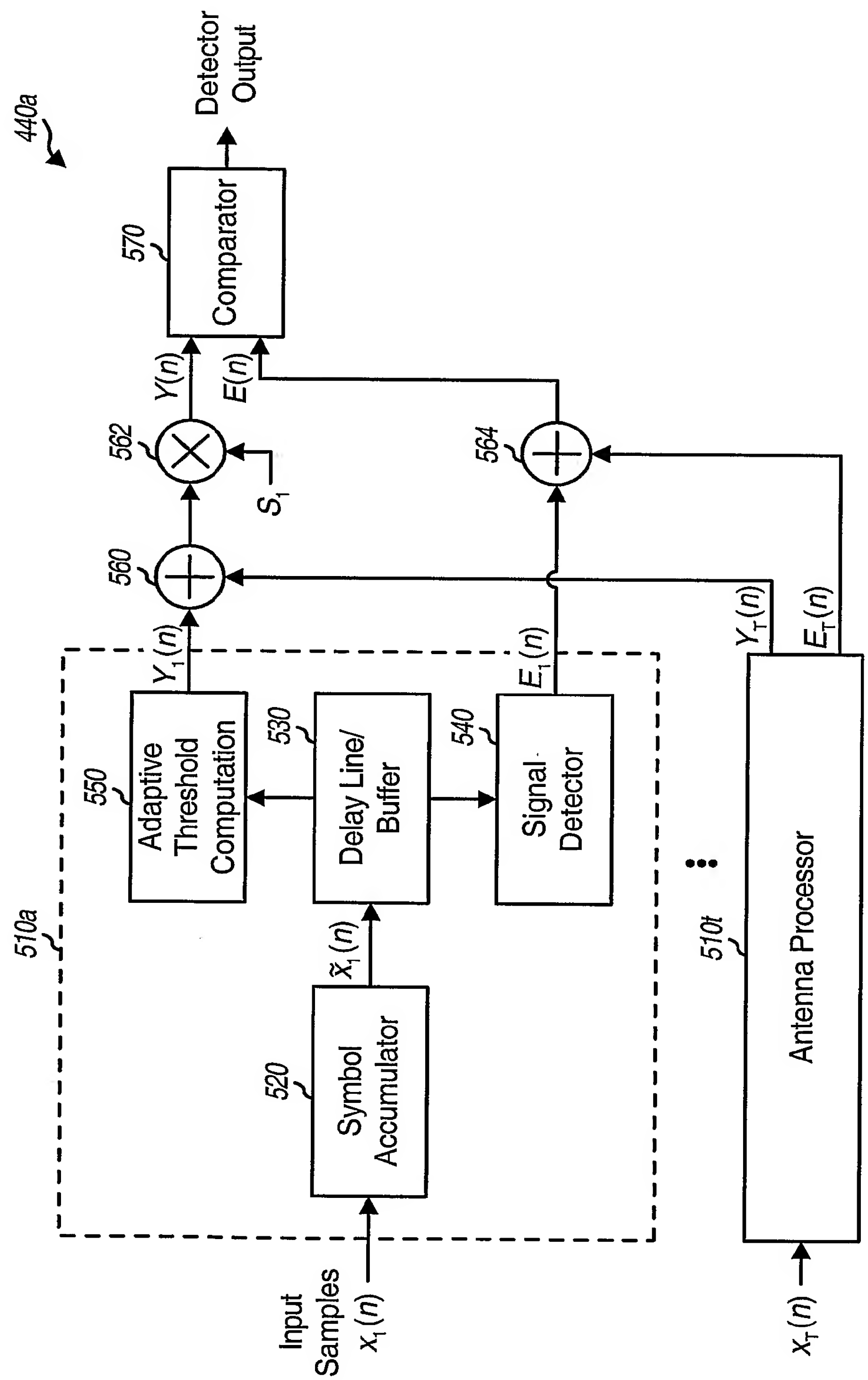

440a
Detector Output
570
Comparator
Y(n)
E(n)
562
S1
560
564
Y1(n)
550
Adaptive Threshold Computation
530
Delay Line/ Buffer
540
Signal Detector
E1(n)
x̃1(n)
520
Symbol Accumulator
510a
Input Samples x1(n)
YT(n)
ET(n)
510t
Antenna Processor
xT(n)

FIG. 5

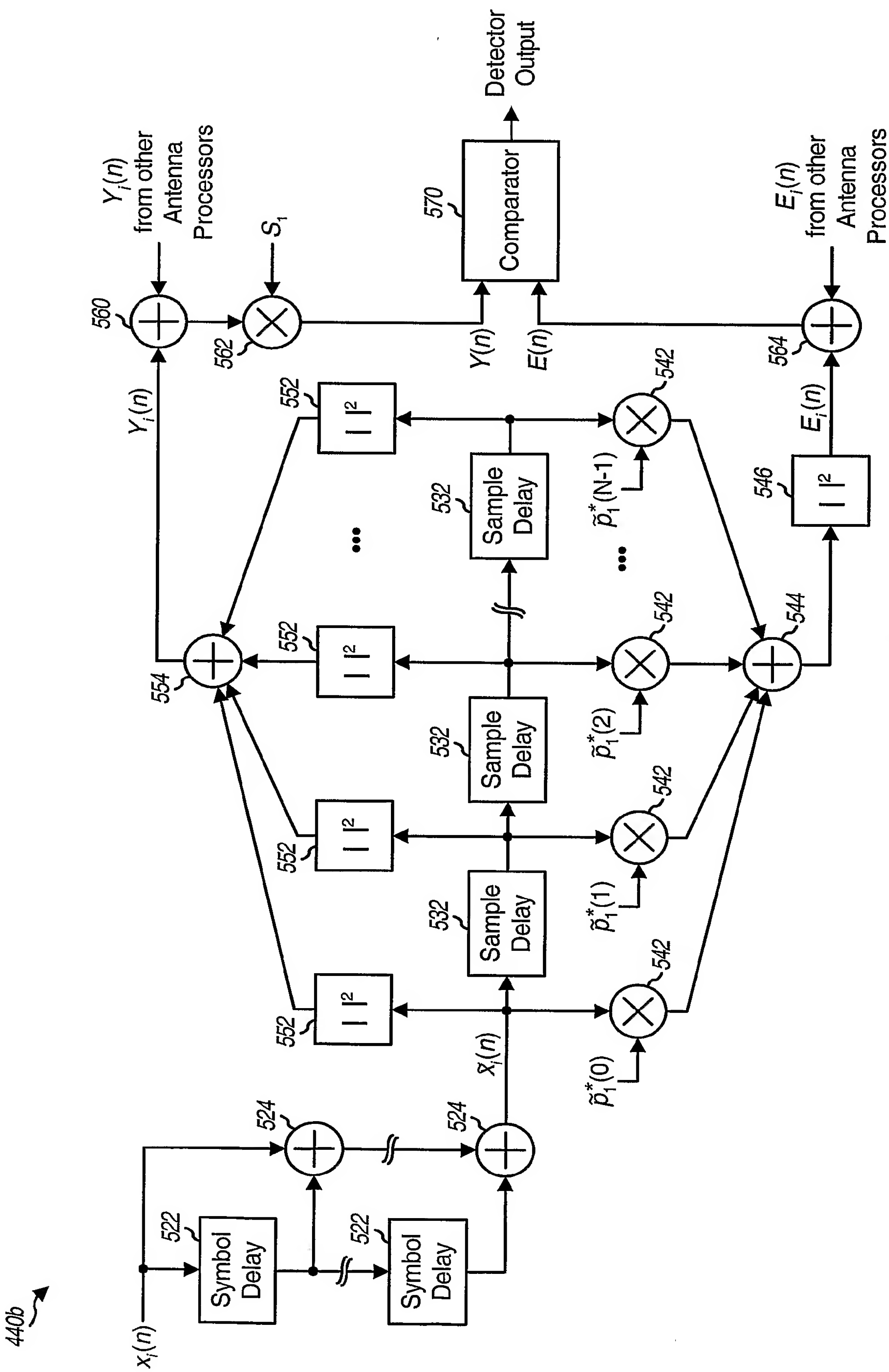
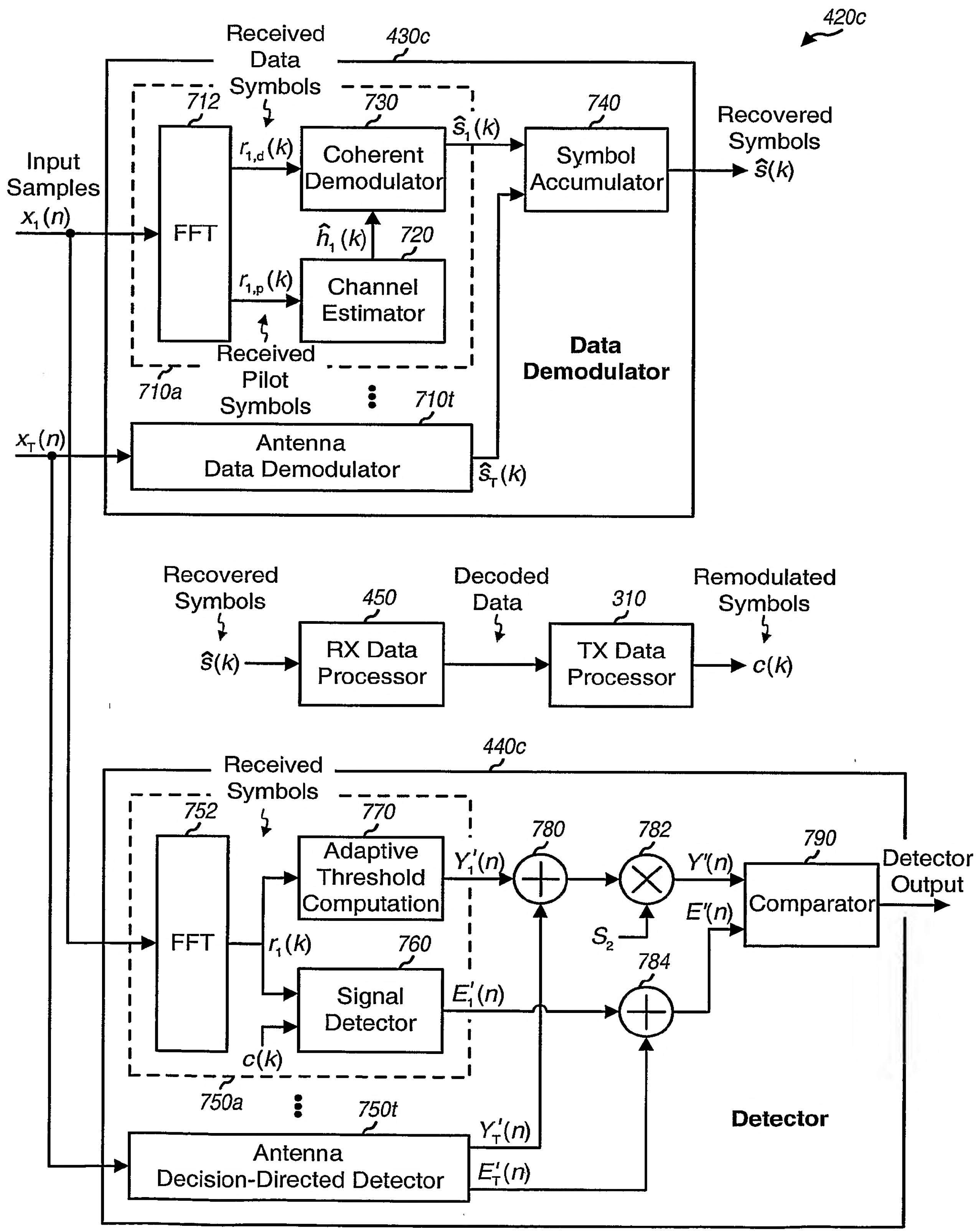

440b
$x_i(n)$
522
Symbol Delay
522
Symbol Delay
524
524
$x_i(n)$
552
$| \, |^2$
532
Sample Delay
542
$\tilde{p}_1^*(0)$
$\tilde{p}_1^*(1)$
$\tilde{p}_1^*(2)$
$\tilde{p}_1^*(N-1)$
554
544
546
$E_i(n)$
$Y_i(n)$
560
$Y_i(n)$ from other Antenna Processors
562
S_1
$Y(n)$
$E(n)$
564
$E_i(n)$ from other Antenna Processors
570
Comparator
Detector Output

FIG. 6



420c
430c
Received Data Symbols
712
730
740
Input Samples
$x_1(n)$
FFT
$r_{1,d}(k)$
Coherent Demodulator
$\hat{s}_1(k)$
Symbol Accumulator
Recovered Symbols
$\hat{s}(k)$
$\hat{h}_1(k)$
720
$r_{1,p}(k)$
Channel Estimator
Received Pilot Symbols
710a
710t
Data Demodulator
$x_T(n)$
Antenna Data Demodulator
$\hat{s}_T(k)$
Recovered Symbols
450
Decoded Data
310
Remodulated Symbols
$\hat{s}(k)$
RX Data Processor
TX Data Processor
$c(k)$
440c
Received Symbols
752
770
780
782
790
Adaptive Threshold Computation
$Y_1'(n)$
$Y'(n)$
Comparator
Detector Output
FFT
$r_1(k)$
760
$E'(n)$
S_2
784
Signal Detector
$E_1'(n)$
$c(k)$
750a
750t
Detector
$Y_T'(n)$
Antenna Decision-Directed Detector
$E_T'(n)$

FIG. 7

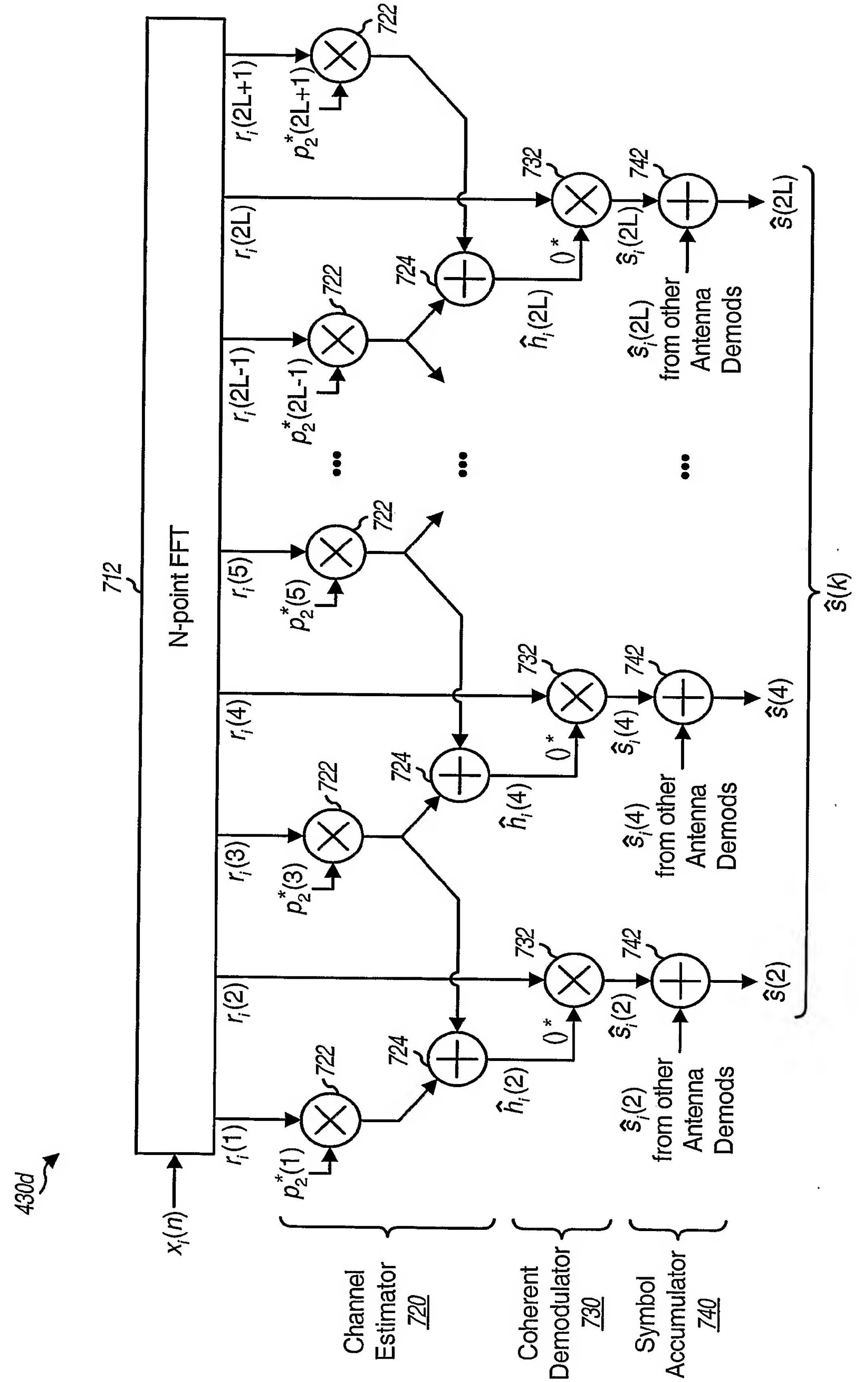

430d
$x_i(n)$
N-point FFT
712
$r_i(1)$
$r_i(2)$
$r_i(3)$
$r_i(4)$
$r_i(5)$
$r_i(2L-1)$
$r_i(2L)$
$r_i(2L+1)$
$p_2^*(1)$
$p_2^*(3)$
$p_2^*(5)$
$p_2^*(2L-1)$
$p_2^*(2L+1)$
722
724
732
742
$\hat{h}_i(2)$
$\hat{h}_i(4)$
$\hat{h}_i(2L)$
$()^*$
$\hat{s}_i(2)$
$\hat{s}_i(4)$
$\hat{s}_i(2L)$
$\hat{s}_i(2)$ from other Antenna Demods
$\hat{s}_i(4)$ from other Antenna Demods
$\hat{s}_i(2L)$ from other Antenna Demods
$\hat{s}(2)$
$\hat{s}(4)$
$\hat{s}(2L)$
$\hat{s}(k)$
Channel Estimator 720
Coherent Demodulator 730
Symbol Accumulator 740

FIG. 8A

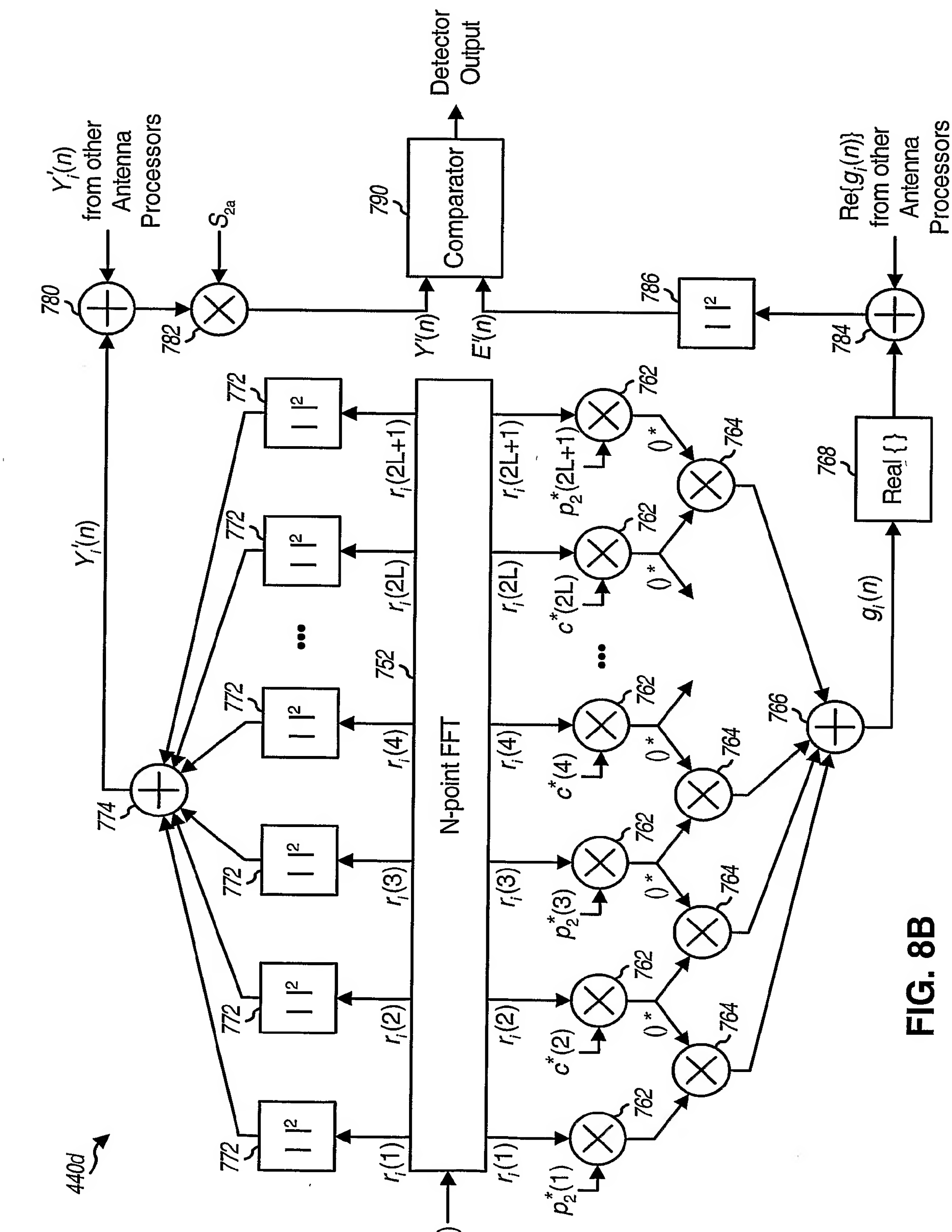

440d
$X_i(n)$
N-point FFT
752
$r_i(1)$
$r_i(2)$
$r_i(3)$
$r_i(4)$
$r_i(2L)$
$r_i(2L+1)$
772
774
$Y_i'(n)$
780
$Y_i'(n)$ from other Antenna Processors
782
S_{2a}
790
Comparator
Detector Output
$Y'(n)$
$E'(n)$
786
784
$Re\{g_i(n)\}$ from other Antenna Processors
$p_2^*(1)$
$c^*(2)$
$p_2^*(3)$
$c^*(4)$
$c^*(2L)$
$p_2^*(2L+1)$
762
764
$()^*$
766
$g_i(n)$
768
Real{ }

FIG. 8B

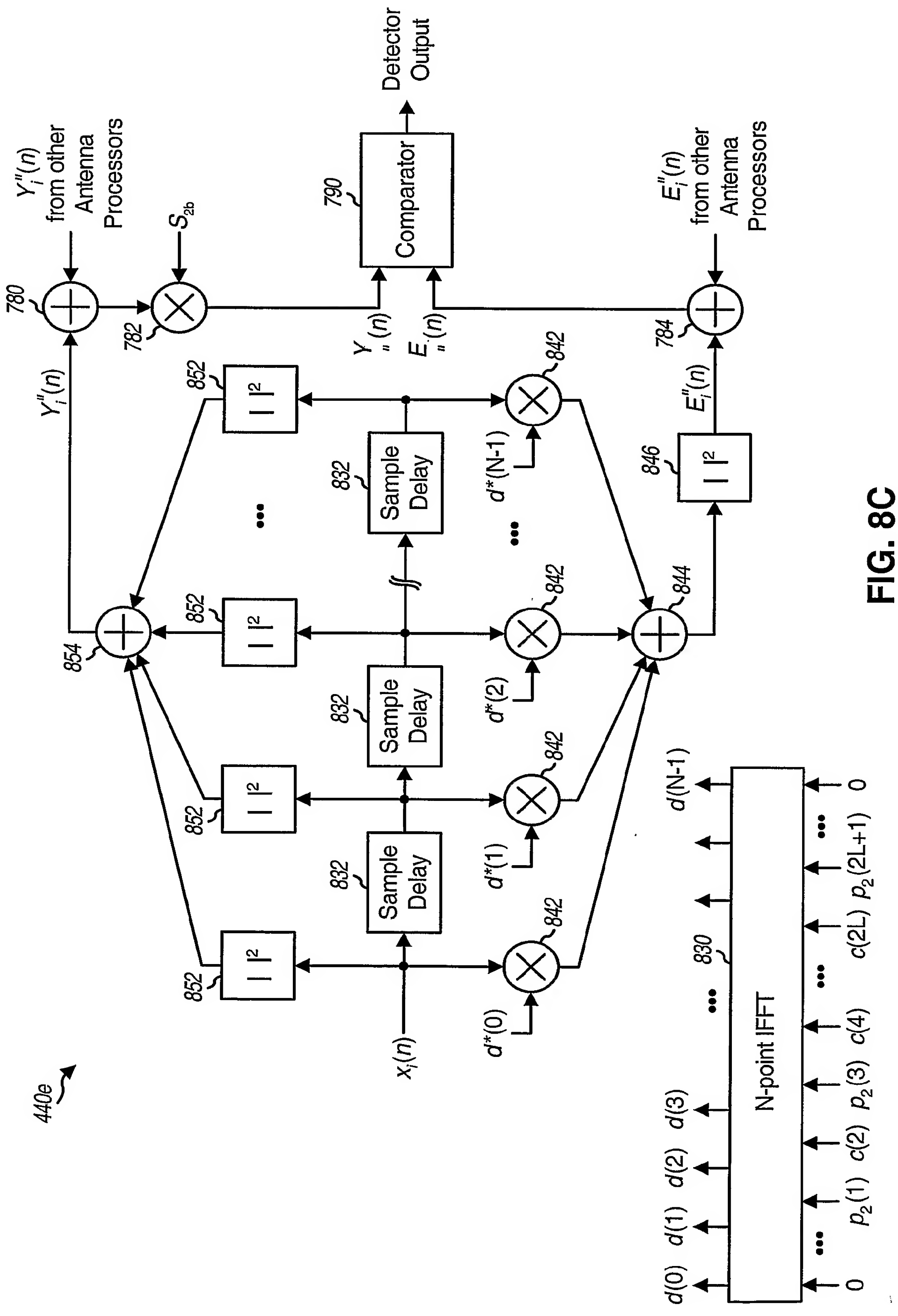

440e
$x_i(n)$
Sample Delay
832
852
842
854
844
846
$d^*(0)$
$d^*(1)$
$d^*(2)$
$d^*(N-1)$
$Y_i''(n)$
$E_i''(n)$
780
782
784
790
S_{2b}
$Y_i''(n)$ from other Antenna Processors
$E_i''(n)$ from other Antenna Processors
$Y''(n)$
$E''(n)$
Comparator
Detector Output
830
N-point IFFT
$d(0)$ $d(1)$ $d(2)$ $d(3)$ $d(N-1)$
0 $p_2(1)$ $c(2)$ $p_2(3)$ $c(4)$ $c(2L)$ $p_2(2L+1)$ 0

FIG. 8C

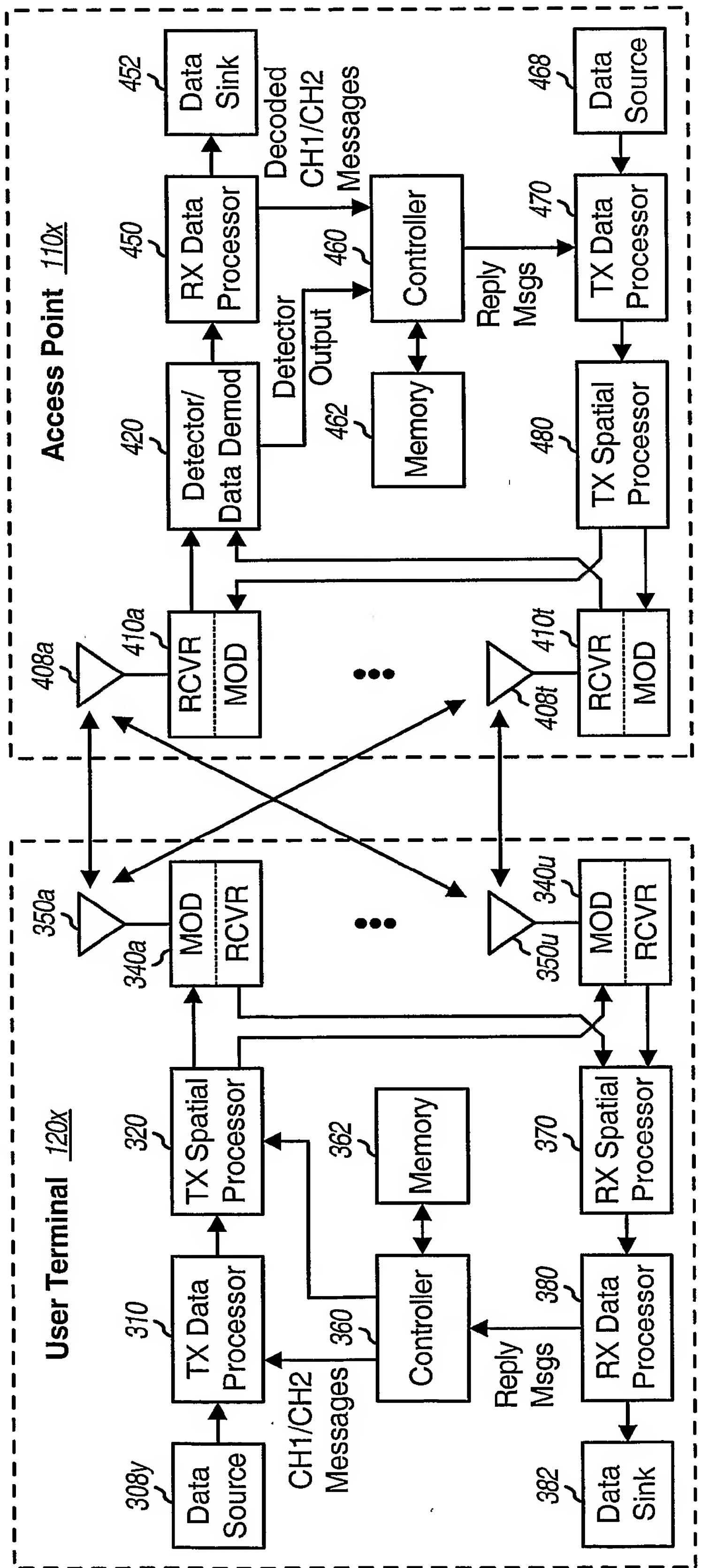

Access Point 110x
408a
410a
RCVR
MOD
420
Detector/
Data Demod
450
RX Data
Processor
452
Data
Sink
Decoded
CH1/CH2
Messages
Detector
Output
460
Controller
462
Memory
Reply
Msgs
470
TX Data
Processor
468
Data
Source
480
TX Spatial
Processor
410t
408t
User Terminal 120x
350a
340a
MOD
RCVR
320
TX Spatial
Processor
310
TX Data
Processor
308y
Data
Source
CH1/CH2
Messages
360
Controller
362
Memory
Reply
Msgs
350u
340u
370
RX Spatial
Processor
380
RX Data
Processor
382
Data
Sink

FIG. 9

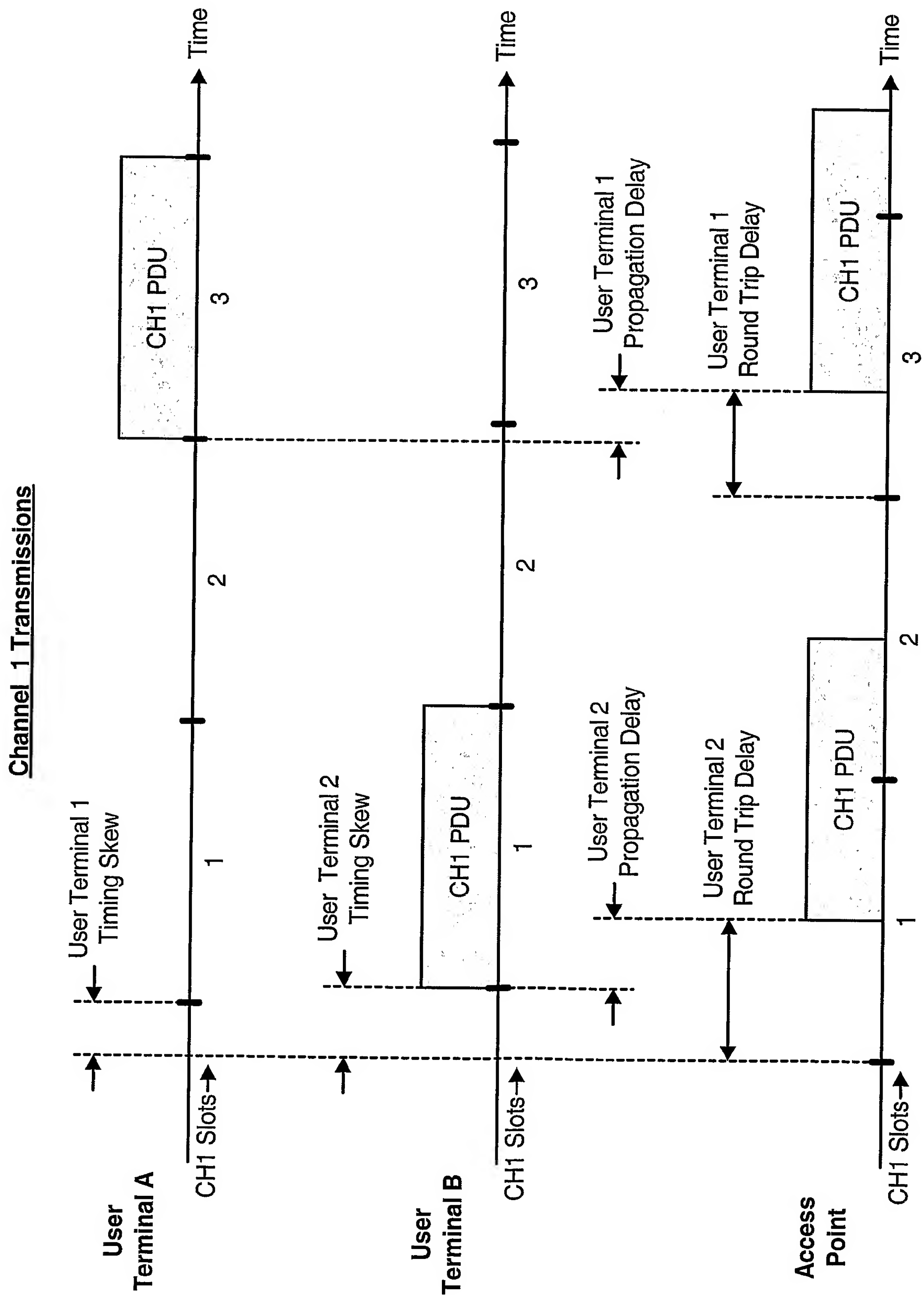

Channel 1 Transmissions
User Terminal A
User Terminal B
Access Point
CH1 Slots
Time
CH1 PDU
1
2
3
User Terminal 1 Timing Skew
User Terminal 2 Timing Skew
User Terminal 1 Propagation Delay
User Terminal 2 Propagation Delay
User Terminal 1 Round Trip Delay
User Terminal 2 Round Trip Delay

FIG. 10A

Channel 2 Transmissions

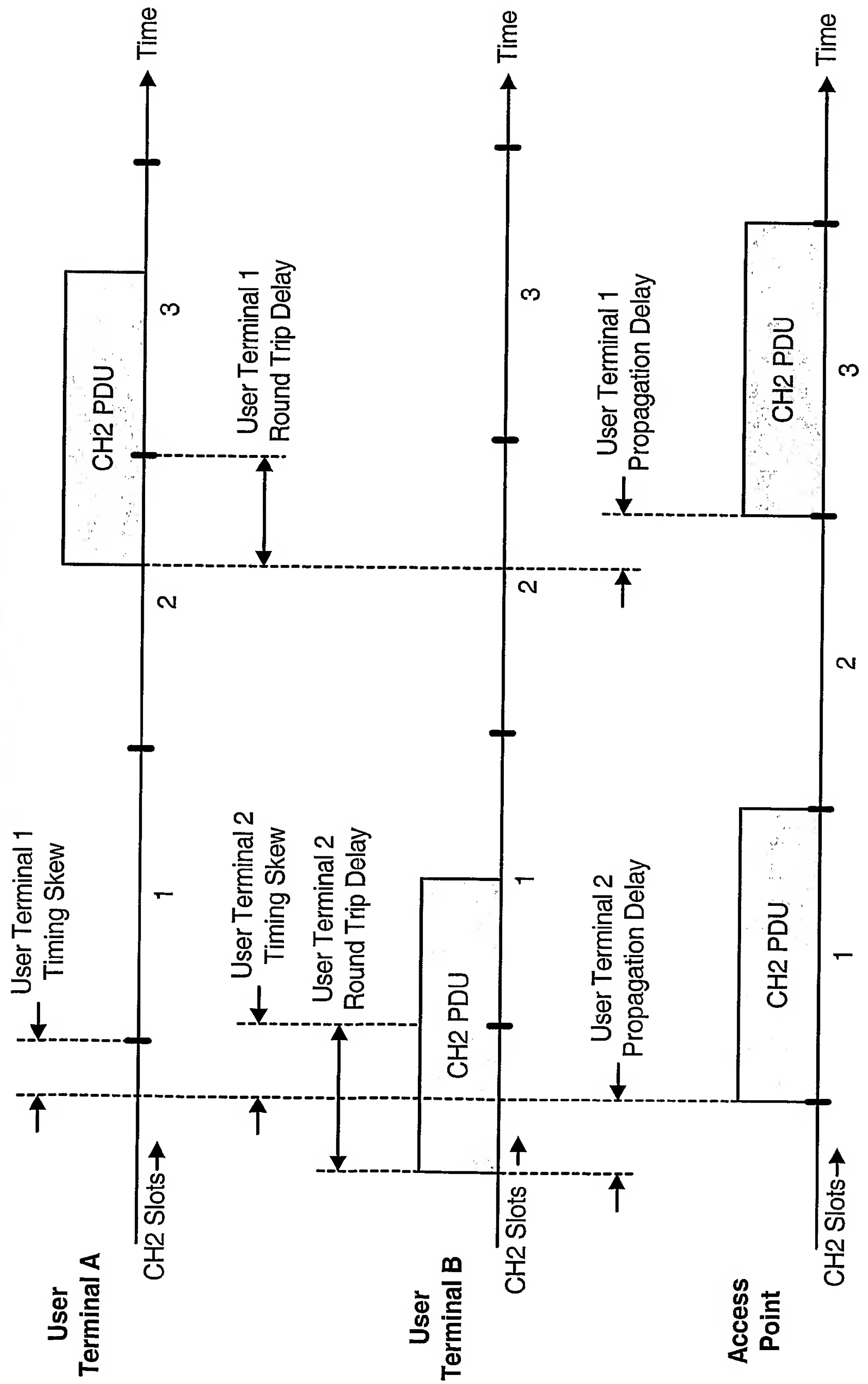


FIG. 10B

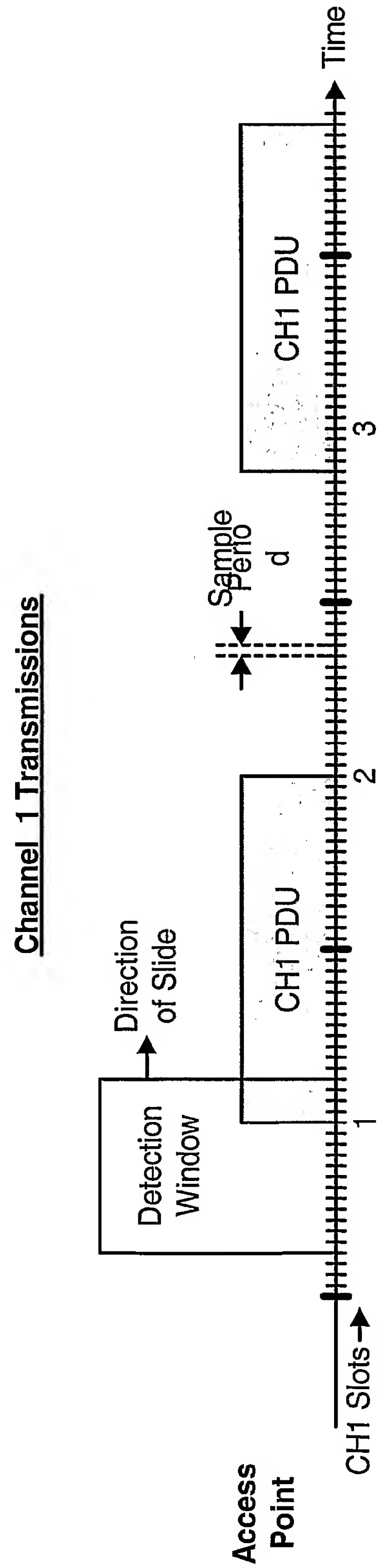


FIG. 11A

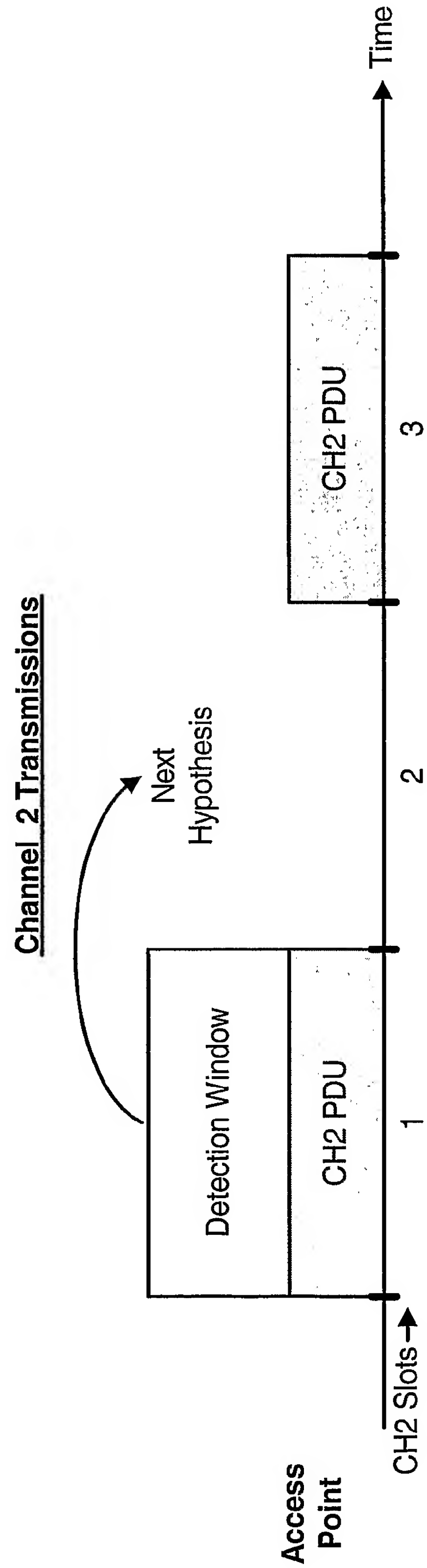


FIG. 11B